

UNIVERSITY OF NOTTINGHAM

Improved receiver tracking models for scintillation monitoring

by

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Abstract

Strong ionospheric electron content gradients may lead to fast and unpredictable fluctuations in the phase and amplitude of the signals from Global Navigation Satellite Systems (GNSSs). This phenomenon, known as scintillation, can impair the tracking performance of a GNSS receiver, leading to increased phase and Doppler errors, cycle slips and sometimes to complete losses of signal lock. In order to mitigate scintillation effects at receiver level, the robustness of the carrier tracking loop, the receiver's weakest link under scintillation, must be enhanced. Thanks to their adaptive nature, Kalman Filter (KF) based tracking algorithms are particularly suitable to cope with the variable working conditions imposed by scintillation. However, the effectiveness of this tracking approach strongly depends on the accuracy of the assumed dynamic model, which can quickly become inaccurate under randomly variable scenarios. This research work shows how inaccurate dynamic models can lead to a KF suboptimum solution or divergence when both strong phase and amplitude scintillation are present. Then, to overcome this issue, two novel self-tuning KF based carrier tracking algorithms are proposed. They self-tune their dynamic models by exploiting the knowledge about scintillation, which is achieved by estimating a number of scintillation indices. These types of tracking schemes are particularly suitable for ionospheric scintillation monitor receivers, which are designed for the computation of scintillation indices and other related parameters. Moreover, this thesis analyses and implements algorithms for a reliable computation of scintillation indices even when low cost receivers are exploited.

Furthermore, a technique is proposed to compute scintillation indices even if temporary losses of signal lock or cycle slips occur. All algorithms have been assessed by exploiting both simulated and real data affected by high latitude and equatorial scintillation. Results show that the proposed algorithms are able to maintain the signal lock and provide reliable scintillation indices when classical architectures and commercial Ionospheric Scintillation Monitoring Receivers (ISMRS) fail.

Preface

This thesis includes materials previously published in conference/journal papers that are referenced below.

M. Susi, M. Andreotti, M. Aquino and A. Dodson (2012) "Ionospheric scintillation effects on the tracking loop of a GPS software receiver," 6th GNSS Vulnerabilities and Solutions Conference, 21-24 May 2012, Baska, Krk Island, Croatia.

M. Susi, M. Andreotti and M. Aquino (2014) "Kalman Filter Based PLL Robust Against Ionospheric Scintillation," Mitigation of Ionospheric Threats to GNSS: an Appraisal of the Scientific and Technological Outputs of the TRANSMIT Project, InTech, DOI: 10.5772/58769.

M. Susi, M. Aquino, R. Romero, F. Dosis and M. Andreotti, (2014) "Design of a robust receiver architecture for scintillation monitoring," 2014 IEEE/ION Position, Location and Navigation Symposium - PLANS 2014, Monterey, CA, 2014, pp. 73-81. doi: 10.1109/PLANS.2014.6851359

R. Romero, **M. Susi**, M. Vuckovic, F. Dosis, M. Andreotti, M. Aquino (2014) "A GPS and Galileo carrier tracking architecture robust to ionospheric scintillation," European Navigation Conference (ENC 2014), Rotterdam, The Netherlands.

M. Susi, M. Andreotti, M. Aquino and A. Dodson (2017) "Tuning a Kalman Filter carrier tracking algorithm in the presence of ionospheric scintillation", GPS Solutions, doi:10.1007/s10291-016-0597-y

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Dedication

To my grandmother Adalgisa and my grandfather Mazzini.

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List of Acronyms

ADC:	Analog Digital Converter
ARSN:	Aviation Radio Navigation Service
AWGN:	Additive White Gaussian Noise
BPF:	Band Pass Filter
CASES	Connected Autonomous Space Environment Sensor
C/N_0 :	Carrier to Noise ratio
DLL:	Delay Lock Loop
FAB:	Fast Adaptive Bandwidth
FFT:	Fast Fourier Transform
FLL:	Frequency Lock Loop
GISM:	Global Ionospheric Scintillation Model
GNSS:	Global Navigation Satellite Systems
GPS:	Global Positioning System
IF:	Intermediate Frequency
ISMR:	Ionospheric Scintillation Monitoring Receiver
KF:	Kalman Filter
LNA:	Low Noise Amplifier
NCO:	Numerical Controlled Oscillator
OCXO:	Oven Controlled Crystal Oscillator
PLL:	Phase Lock Loop
PRN:	Pseudo Random Noise
PSD:	Power Spectral Density
PVT:	Position Velocity Time
RHCP:	Right Hand Circularly Polarized
RF:	Radio Frequency
RNSS:	Radio Navigation Satellite Service
SAKF:	Scintillation Based Adaptive Kalman Filter

SI:	Scintillation Intensity
SNV:	Signal to Noise Variance
SPLN:	St Petersburg, and Leeds/Newcastle University
TEC:	Total Electron Content
TCXO:	Temperature Compensated Crystal Oscillator
USRP	Universal Software Radio Pheriperal
WBMOD	Wide Band Model

Chapter One: INTRODUCTION

Global Navigation Satellite Systems (GNSS) currently play a key role in modern society enabling applications in several crucial sectors. Strategic areas of applications include vehicular and personal navigation, aircraft and maritime navigation, location based and rescue services. However, despite its worldwide success and diffusion, GNSS is still a sensitive system vulnerable to failure and disruptions. This is of particular concern for users of safety of life services demanding high reliability, availability and continuity of service. The disruptions potentially threatening GNSS are usually classified as intentional and unintentional. Intentional disruptions, such as jamming and spoofing, are produced to deliberately impair GNSS receiver operation. Unintentional disruptions can include man-made interference, for example originating from satellite communications, TV broadcasting and Ultra Wide Band (UWB) communications, and natural interference, due to space weather related events. One of the main natural threats to the reliability and availability of GNSS is represented by the non-stationary propagation conditions experienced by Radio Frequency (RF) signals when traversing the ionosphere. In particular small-scale ionospheric irregular structures may refract and diffract (Kirtner et al., 2009) GNSS signals producing random and fast variations in their amplitude and phase (Rino et al., 1979), an effect known as scintillation. Amplitude scintillation manifests itself as instantaneous increases and decreases of the trans-ionospheric signal intensity. This phenomenon, when severe, can lead to deep signal fading and, consequently, induce the signal to noise ratio to drop below the receiver tracking threshold. Moreover, phase

scintillation could increase the Doppler shift and render it larger than the phase lock loop bandwidth. As a consequence, cycle slips or even a loss of lock could occur. Even though this phenomenon does not usually affect all satellites in view at the same time, involving only a portion of the sky, it may be able to degrade the final position solution accuracy. Moreover, if the unaffected satellite links are not sufficient to provide a solution, outages in the GNSS operation could be experienced. A way to mitigate scintillation effects on GNSS is to increase the robustness of GNSS receivers, and in particular that of the carrier tracking stage, which is the part of the receiver most vulnerable to scintillation effects (Hinks et al., 2008).

1.1 Robust carrier tracking under scintillation: general overview

Conventionally, the carrier phase is obtained by a closed tracking loop, the Phase Locked Loop (PLL), which extracts the carrier phase measurements from the noisy input. For this purpose, as detailed in Chapter 2, a local signal replica, matching the input signal as much as possible, is generated by the Numerical Controlled Oscillator (NCO) and correlated with the input signal (Kaplan and Hegarty, 2006). The correlator outputs are then input to the discriminator, a non-linear device estimating the error in the parameters to be tracked (O'Driscoll et al, 2011). Then the tracking error is filtered by the loop filter and exploited to adjust the NCO which drives the generation of a new signal replica so as to minimize the tracking error.

The robustness of the PLL is determined by the capability to also maintain the signal lock in non-nominal conditions, as, for instance, in the presence of weak signals or high dynamics produced by scintillation. Difficult scenarios are characterized by weak signals

and/or fast signal dynamics, which require a smart PLL design. To cope with weak signals narrow PLL bandwidths, or longer time of integration values, are usually exploited. To deal with fast signal dynamics, wide loop bandwidths, or shorter time of integration values, should be preferred (Kaplan and Hegarty, 2006).

In Figure 1.1 the trade-off between noise reduction, due, for instance, to amplitude scintillation and thermal noise, and the agility in following the signal dynamics, i.e. produced by phase scintillation, clock noise, and receiver dynamics is graphically illustrated for the PLL filter bandwidth and time of integration.

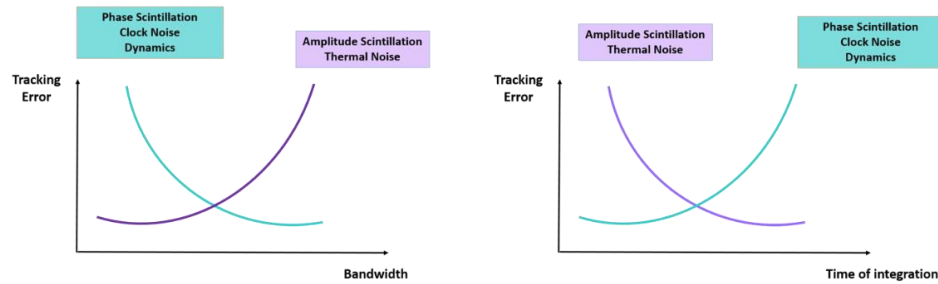


Figure 1-1: Tracking error versus bandwidth (left) and time of integration (right).

Such constraints are particularly challenging for the design of a receiver robust under scintillation and especially in the case of equatorial scintillation. Indeed the latter is the most critical scenario, since it is characterized by canonical fades, a term used by Hinks et al. (2008) to define simultaneous deep signal fading (up to 25 dB) and abrupt phase variations typical of equatorial scenarios.

A very good review on carrier tracking architectures that are robust under critical scenarios can be found in Lopez-Salcedo et al. (2014). An effective way to avoid the

design dilemma typical of closed loop tracking architectures is to use open loop architectures (i.e. Curran et al., 2014). The latter are based on a feed forward estimation of signal parameters by processing a batch of signal parameters each time. These architectures do not rely on feedback information and, consequently, they do not lose the signal lock, ensuring an increased robustness. However, the main drawback of these algorithms is their lower accuracy and high complexity. Another possible approach is to replace the PLL with an FLL which is more robust but also less accurate. When high accuracy is required, the FLL can be used as a backup solution to replace the PLL in case of deep signal fades (e.g. Fantinato et al., 2012). Alternatively, FLL assisted PLL techniques can be employed (Xu et al., 2015, Chiou et al. 2007). They allow estimating the frequency and phase errors, which are then combined to adjust the NCO. Furthermore, adaptive tracking schemes (e.g. Skone et al., 2005, Tiwari et al. 2011) represent a suitable solution to optimize the tracking parameters in the presence of variable conditions. They can allow tuning the loop filter bandwidth according to an optimization criteria, which may depend on different parameters, such as the estimated Carrier to Noise Ratio (C/N_0) and the signal dynamics.

Among the adaptive techniques the Kalman filtering represents an appealing solution to cope with scintillation induced fast dynamics and deep fading (i.e. Macabiau et al., 2012; Psiaki et al., 2007). However, in order to ensure the optimization of the KF, some assumptions should be fulfilled. First of all, the additive noise should be white and Gaussian, then the process noise covariance and the measurement noise should be known (Brown and Hwang, 1997). The robustness of a GNSS receiver under scintillation can also be enhanced by exploiting multi-frequency and multi-constellation solutions.

These architectures track all signals simultaneously, combining the tracking and navigation stages. In this way the user and satellite positions can be exploited to estimate phase and code measurements (Peng, 2012). Moreover, the satellite links with stronger signals can be used to aid the ones with weaker signals. This aspect is particularly attractive in the case of scintillation, when usually only some satellite links are affected. Furthermore, since scintillation affects different frequencies in different ways (lower frequencies are more heavily affected) tracking loops combining the multi-frequency signals could be exploited under scintillation. An example of multi-frequency and multi-satellite aiding architecture has been described by Lin et al. (2014). In this architecture, considering that the GPS L1 C/A signal is more robust to scintillation than the GPS L2C, the GPS L2C PLLs are aided by the GPS L1 C/A PLL. These multi-frequency and multi-satellite aiding algorithms usually exploit KFs as local filters. More details about these tracking architectures that are robust under harsh scenarios will be given in Chapter 5.

1.2 Limitations of previous works and objectives

Although in the literature several tracking schemes have been proposed to improve the resistance of GNSS receivers under harsh tracking scenarios, KF based tracking architectures have emerged as one of the most attractive solutions under critical working conditions, i.e. in presence of deep fading or fast signal dynamics. Moreover, as underlined before, the KF is at the core of promising architectures based on multi-frequencies and multi-satellites, which will become more successful with the full deployment of new constellations.

The effectiveness of KFs under difficult scenarios, can be attributed to their intrinsic adaptive nature. Indeed the KF automatically adjusts its equivalent bandwidth to minimize the mean square error of the estimated parameters (Brown and Hwang, 1997). However, this approach is optimum when the measurement noise and dynamic models are known a priori, as pointed out above. This aspect can be particularly challenging under scintillation scenarios, where the working conditions can rapidly change, making the a priori assumed dynamic and measurement models inaccurate. Regarding this concern, little work has been done in the literature to specifically optimize the KF for scintillation scenarios.

For this reason this research work aims to design and implement tracking algorithms specifically designed to cope with scintillation scenarios. Indeed, although it is not likely that a mass-market receiver would be optimized under scintillation, this could be advantageous for Ionospheric Scintillation Monitoring Receivers (ISMRs), which have gained great success and diffusion in the last few years. ISMRs provide information on ionospheric disturbances through the estimation of a number of ionospheric related parameters, including indices that are commonly used to quantify the level of scintillation (Bougard et al., 2011). These receivers, which are usually placed in locations where scintillation is likely to occur, require an enhanced robustness under scintillation since poor tracking performance will translate into impaired scintillation monitoring capabilities. In order to provide an accurate solution, commercial ISMRs are usually geodetic high grade receivers, which also rely on the carrier phase information.

In this thesis it is shown that the information provided by ISMRs on the working conditions, i.e. on the scintillation level, could be exploited to adjust the tracking scheme

in an optimum way in order to enhance its robustness. The proposed techniques could also be exploited for low cost receivers, which do not include high quality oscillators and that were not originally designed to monitor scintillation. As shown in this dissertation, in this case some signal processing techniques should be applied to reduce the phase noise introduced by the low quality clock. All the techniques proposed here have been implemented into a software receiver that, thanks to its configurability and flexibility, represents a powerful and low cost tool to implement, assess and compare different tracking architectures. It should be underlined that although these tracking schemes have been assessed based on the tracking of GPS L1 signals, the same approach could be used to track other signals.

The main objectives of this research work can be summarized as follows.

- 1. Analyse the effects of front end and clock characteristics on the scintillation monitoring capabilities of a GPS receiver**

The algorithms proposed here are mainly designed for ISMRs, which by definition are designed to provide scintillation indices and related parameters. A reliable computation of these scintillation parameters is also fundamental for the effectiveness of the tracking algorithms proposed in this thesis. However, the computation of these parameters can be affected by the low quality clock and hardware features (Datta-Barua et al, 2015) in the case of low cost receivers. To overcome this issue specific signal processing techniques to remove external noise sources other than from scintillation have been analysed and implemented.

- 2. Implement algorithms for reliable scintillation monitoring**

Different algorithms to estimate scintillation parameters have been implemented and their performance compared with that of commercial ISMRs. Specifically, the algorithms allow estimating the amplitude and phase scintillation indices, respectively S_4 and Φ_{60} , along with the spectral strength of the phase noise at 1 Hz and the spectral slope of the phase Power Spectral Density (PSD), respectively T and p . It has been observed that in the case of commercial ISMRs a main issue arises in the computation of the phase related scintillation indices (Φ_{60} , p , T) when temporary losses of lock or cycle slips occur. Indeed, the latter induce the phase detrending filter, exploited to estimate these parameters, to reset, producing, as a consequence, data gaps in the estimation of these parameters. To overcome this issue, sliding windows are exploited for the computation of Φ_{60} in order to increase the available samples, minimizing the probability of data gaps occurring. This approach allows obtaining continuous phase indices information when commercial ISMRs fail.

3. Investigate the effects of an inaccurate KF tuning under scintillation

As previously mentioned, the literature has shown that KF based tracking is particularly suitable under scintillation scenarios. However, the effectiveness of the KF based tracking scheme is closely related to the accuracy of the dynamic/measurement model, which usually is defined a priori. This research shows that when the working conditions quickly change over time, as in the case of scintillation, the initially assumed dynamic model may not be valid anymore, leading to a sub-optimum solution or, in the worst cases, to a filter divergence.

4. Propose algorithms to tune a KF based tracking scheme in the presence of scintillation

In order to overcome the issues of poor KF tuning, pointed out above, this thesis proposes two novel self-tuning KF based tracking schemes. The first tracking algorithm, indicated as a Scintillation based Adaptive Kalman Filter 1 (SAKF1), uses the scintillation spectrum parameters p and T , continuously estimated, to include the scintillation phase error contribution in its covariance matrix. The second algorithm, indicated as Scintillation based Adaptive Kalman Filter 2 (SAKF2), has been proposed to reduce the computation cost of SAKF1. The latter requires only the phase scintillation index Φ_{60} evaluation, which is usually performed by any scintillation monitoring receiver. This algorithm first detects the level of phase scintillation and then selects the appropriate dynamic model among four pre-defined ones. Both algorithms also exploit a scaling factor depending on the filter residuals to weight the measurement noise. Furthermore, the above tracking architectures, implemented into a software receiver, have been assessed by using simulated and real data affected by both high latitude and equatorial scintillation. Both the tracking schemes also include an algorithm for the computation of scintillation indices. Results show that these algorithms outperform conventional techniques and commercial ISMRs.

5. Investigate models to reproduce GNSS signals affected by scintillation

In order to assess the proposed algorithms in terms of tracking robustness and scintillation monitoring capabilities, different scintillation models exploited to reproduce GNSS signals affected by scintillation have been investigated. Since real data has been made available only for the equatorial region, particular emphasis has been given to scintillation models able to reproduce high latitude scintillation.

6. Investigate the different effects of high latitude and equatorial scintillation on the receiver tracking loop

As previously mentioned, the performance of the proposed algorithms has been compared to that of conventional tracking schemes and commercial ISMRs under both high latitude and equatorial scintillation. This assessment phase offered the opportunity to better understand and analyse the different effects of high latitude and equatorial scintillation on the GNSS receiver tracking stage.

1.3 Thesis organization

The structure of this dissertation is summarized as follows.

Chapter 2 reviews the theoretical background of GNSS. It provides an overview of the main error sources affecting GNSS, and then gives details of the structure of GPS signals. Finally, it introduces the general structure of a GPS receiver with particular emphasis on the receiver tracking stage, which is the main focus of this thesis.

Chapter 3 provides an overview on the ionosphere and its main effects on GNSS signal propagation. Particular emphasis is given to ionospheric scintillation. The chapter covers the physical processes behind ionospheric scintillation underlining the differences between high latitude and equatorial scintillation. Finally, the chapter introduces several scintillation models, which can be exploited to reproduce both high latitude and equatorial scintillation.

Chapter 4 reviews the main parameters used to quantify the level of scintillation. Furthermore, this chapter details the different techniques to compute these scintillation

indices by using a software receiver. Moreover, the signal processing techniques exploited to compute scintillation indices using low cost receivers are also detailed.

Chapter 5 introduces the main features required for a scintillation monitoring receiver. It then analyses in depth several tracking architectures that are robust under harsh scenarios such as under scintillation. Focus is given to KF based tracking algorithms. Moreover, the two novel KF based tracking schemes, indicated as SAKF1 and SAKF2 herein, are proposed and detailed.

Chapter 6 is dedicated to the assessment stage of the algorithms proposed in this thesis. First, the experimental set-up for the data collection is described. Finally, the performance of the algorithms is compared with that of conventional tracking schemes and commercial ISMRs. The comparison is carried out in terms of tracking and scintillation monitoring performance.

Finally **Chapter 7** draws some conclusions and proposes research directions for future work.

Figure 1-2 shows a scheme representing the thesis's structure, along with the logic interconnections among chapters and the general topic covered by each chapter.

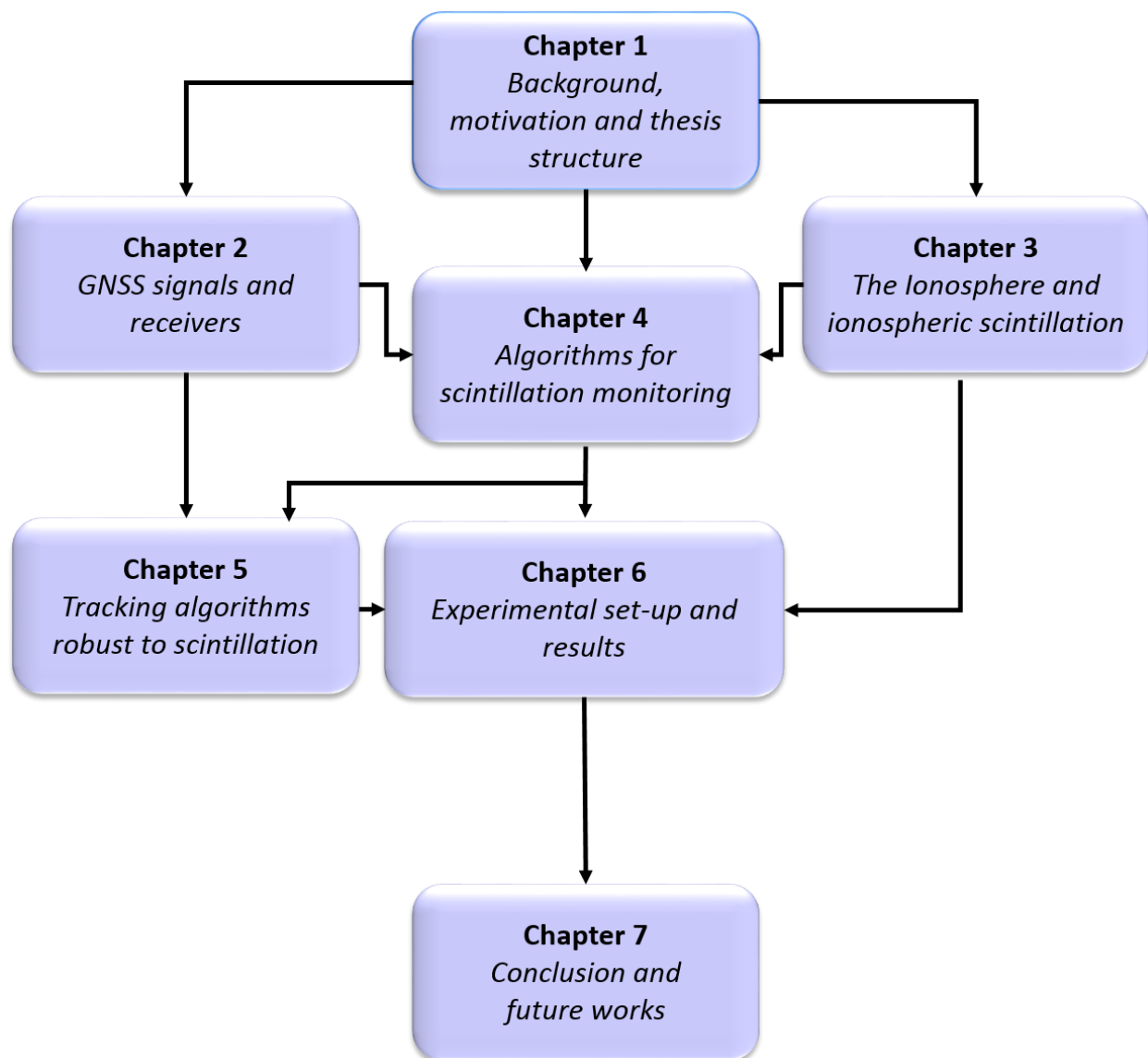


Figure 1-2: Structure of the thesis and logic interconnections among chapters.

Chapter Two: GNSS Background and Receiver Structure

This chapter provides a general overview of Global Navigation Satellite Systems (GNSS) as background to the understanding of the next chapters of this dissertation. Specifically the chapter first describes the general working method and structure of a GNSS; then it gives details on GNSS signals with particular focus on GPS L1 C/A signals, which are used to assess the algorithms proposed in this thesis. Furthermore, the chapter illustrates the various blocks composing a GNSS receiver and their functionalities. Particular emphasis is placed on the receiver's tracking stage, the design of which is the main focus of this thesis. Finally, the main error sources affecting the performance of a GNSS receiver are briefly described.

2.1 GNSS: general overview

The etymology of the word navigation comes from the Latin noun *navis* (*ship*) and the verb *agare* (*to drive*). The term was introduced in the 15th century during the age of discovery to indicate the art of driving a ship from a starting point to a destination by estimating its position and direction. However, the act of navigating is much more ancient. Already many years before the 15th century, around 4000 years ago, pioneering sailors such as the Phoenicians used to traverse the Mediterranean by relying on the indications of primitive maps and using the stars as celestial markers. Many years later, in the 19th century, RF signals were exploited to estimate the user position through the so termed radio navigation techniques. In modern times the term navigation was then used to indicate the art of

determining the position and velocity of any moving object over time and did not refer only to ships. Position and navigation by means of electronic systems are traditionally accomplished by following two main approaches, indicated as active and passive ranging (Kaplan and Hegarty, 2006). Active ranging systems, such as radar or sonar, are based on the transmission of signals to target objects that reflect back the received signals. Then the computation of the distance between transmitting and target devices can be estimated through the evaluation of the round trip delay and indeed they are defined as two way ranging systems.

GNSSs belong to the category of passive ranging systems where the targets are indicated as passive devices since they only receive signals without needing transmission capabilities. For this reason, these types of systems are indicated as one way ranging systems. A GNSS is a worldwide system that provides position, velocity and time information by exploiting satellite signals. The Global Positioning Satellite System (GPS), which was developed by the United States Department of Defence (DoD) in 1973, is the first and most common GNSS system. The GPS was developed in order to give precise position and time information for military applications, but over the years it was also extended to civilian applications. At the time of the writing of this dissertation only GPS and the Russian GLObal NAVigation Satellite System (GLONASS) are fully in operation while the Chinese and the European GNSSs, respectively indicated as Beidou and Galileo, are under development. All GNSSs share the same type of structure composed of a space, a ground control and a user segment. A general overview of this structure is presented in the following section.

2.2 GNSS Structure

A GNSS is composed of three main segments. They are the Space Segment (SS), the Ground Control Segment (GCS) and the User Segment (US). The SS is a constellation of artificial satellites orbiting the Earth and transmitting the GNSS signals to both the GCS and the US. GNSS satellites receive the navigation message from the GCS and rebroadcast it to the users. Characteristics of the broadcasted GNSS satellites are detailed in Section 2.4. Satellites house very stable atomic clocks, which provide to the US the accurate time of signal transmission necessary for the computation of the Position Velocity Time (PVT) solution (Kaplan and Hegarty, 2006). A generic GCS is composed of a network of monitor stations, one or more control stations and several uplink stations. The monitor stations, which are located at precise surveyed locations, send the ranging measurements from the satellites to the control stations. The monitor stations house synchronized clocks so that their ranging measurements can be exploited to estimate satellite orbits and calibrate satellite clocks. To summarize, the main CGS functionalities are (Kaplan and Hegarty, 2006) as follows:

- To monitor satellite orbits
- To check and maintain satellite health
- To maintain GNSS time
- To predict ephemeris and clock parameters
- To update satellite navigation messages

The GCS network for the case of the GPS is shown in Figure 2-1 where it can be seen the Master Control Station placed at Schriever Air Force Base (AFB) in Colorado, an alternate master control station, 12 command and control antennas and 16 monitoring stations.

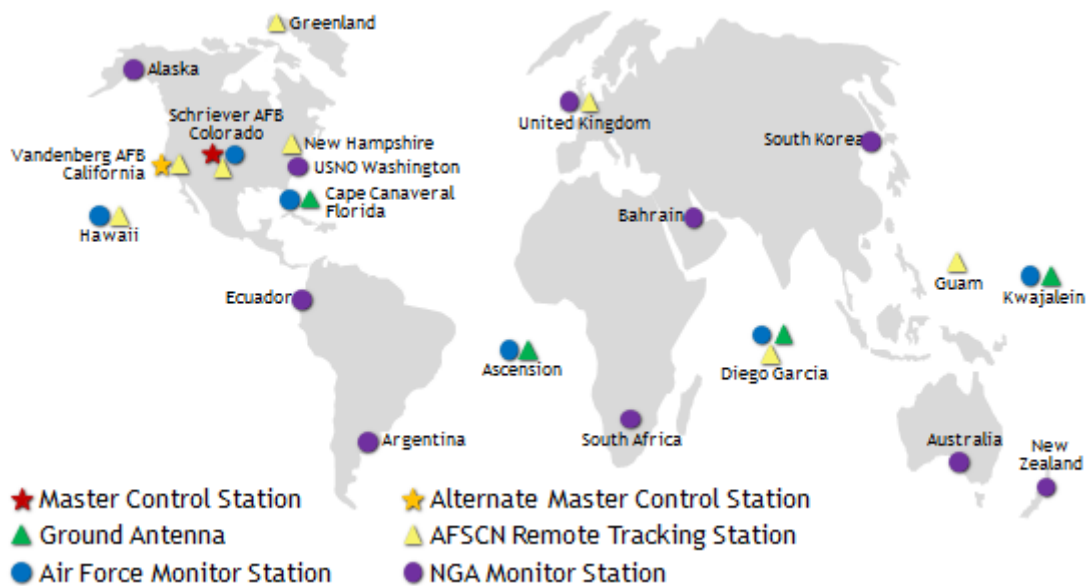


Figure 2-1: GPS GCS representation (source: <http://www.gps.gov/systems/gps/control/>).

The US is composed of all user receivers, which receipt the GNSS signals and process them to compute the final solution. The features and capabilities of these receivers change according to the application and user requirements. As detailed in Section 2.5, despite their different characteristics any receiver has the common objective to check all satellites in view, estimate the distance between satellites and user and, finally, to compute the Position Velocity and Time (PVT) solution.

2.3 GNSS signals: general overview

GNSSs are based on the transmission of RF signals. Conventionally, they exploit a Direct Sequence Spread Spectrum (DSSS) modulation (Ziemer et al., 1995), which consists of multiplying the data message by a Pseudo Random Noise (PRN), usually a binary sequence, termed as the spreading code occupying a bandwidth much wider than the signal data. Since a unique PRN code is transmitted from each satellite, it is possible to transmit more signals at the same frequency and univocally identify each broadcasting satellite (Pany, 2010). The data message includes precise orbital parameters, namely ephemeris information, which enables to compute the exact satellite position at the time of the transmitting signal. The data also embeds coarse information about the other satellites, indicated as almanac.

Therefore, the receiver can obtain the ephemeris and almanac information by demodulating the above data. The spreading code is synchronized to the satellite clock, and all the satellite clocks are synchronized to the GNSS time thanks to the GCS. The signal transmitted by one satellite takes a time τ to arrive at the user receiver. This time can be computed by the receiver by exploiting the time of signal transmission embedded into the navigation message. By receiving signals from at least three satellites, theoretically, the receiver should be able to compute its position by applying the trilateration concept. Indeed if the exact position of three satellites is known, along with the distances between the receiver and the above satellites, the receiver position can be geometrically determined. A minimum of three satellites are necessary because of the three unknown variables, namely latitude, longitude and height, which are necessary to

localize the receiver. By multiplying the Time of Arrival (TOA) ranging by the velocity of light ($c = 299,792,458 \text{ m/s}$) the distance between satellite and user can be computed (Misra and Enge, 2011). However, to apply a pure trilateration process the exact time of signal transmission and arrival should be known and measured on the same time scale. Satellites house very expensive and accurate atomic clocks which are synchronized to a common time scale; in contrast, the clocks included in the user receivers, for cost reasons, are usually less accurate.

Consequently, since the receiver clock is not synchronized with the satellite clock, the measured signal time of transmission is a biased version of the real one and the measured range is not the real one. Therefore, the measured range is indicated as pseudorange to underline its difference with the real range. Also, the clock bias is considered as a further unknown and its effect has to be removed. For this reason, as illustrated in Figure 2-2, at least four satellites are necessary to solve the system including four unknown parameters (i.e. latitude, longitude, height, time bias) (Kaplan and Hegarty, 2006). Moreover, it should be noted that other error sources could affect the position estimated, as detailed in Section 2.6.

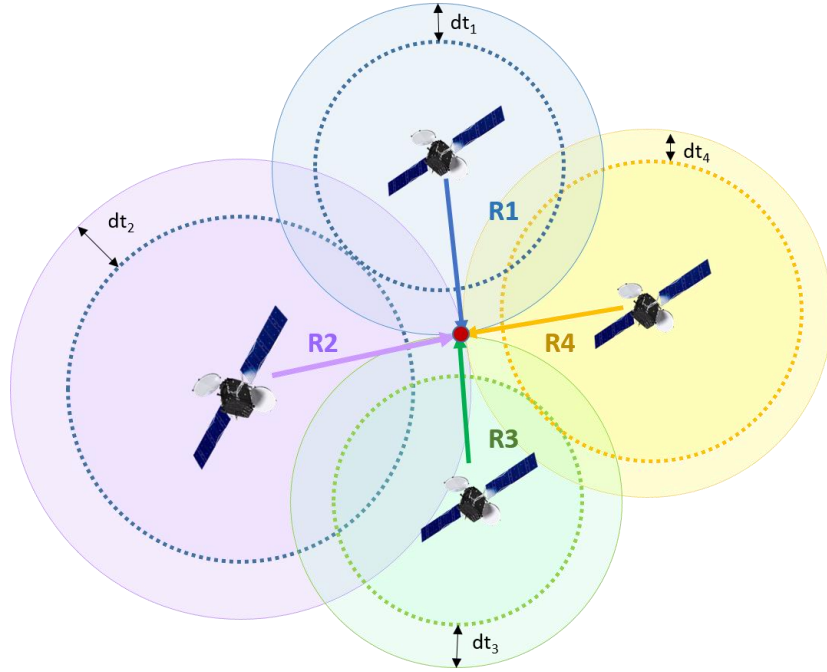


Figure 2-2: Graphical representation of the trilateration process when the receiver clock is biased.

2.4 Signal model

A generic GNSS signal arriving at the input of a GNSS receiver antenna can be modelled as (Misra and Enge, 2011):

$$r_{RF}(t) = \sum_{i=1}^K s_{RF}^i(t) + n(t) \quad (2.1)$$

Where $r_{RF}(t)$ is given by the sum of the K useful component $s_{RF}^i(t)$ transmitted by the K satellites in view plus a noise component $n(t)$. The noise component is modelled as a zero-mean Additive White Gaussian Noise (AWGN) process with a two sided Power Spectral Density (PSD) of $N_0/2$ [W/Hz]. This term is produced by any noise source affecting

the signal propagation channel. Each useful signal component $s_{RF}^i(t)$ can be further expanded as:

$$s_{RF}^i(t) = \sqrt{2P_i(t)}d_i(t - \tau_i)c_i(t - \tau_i)\cos\left(2\pi(f_{RF} + f_D^i)t + \varphi_i(t)\right) \quad (2.2)$$

Where

- $P_i(t)$ is the power of the i^{th} signal arriving at the GNSS receiver antenna.
- $d_i(t)$ is the bipolar navigation message assuming values in the set $\{-1, 1\}$. It includes the satellite health status, ephemeris, clock bias parameters and an almanac with coarse ephemeris data of all the other satellites in the constellation.
- τ_i represents the delay introduced by the communication channel.
- $c_i(t)$ is the spreading code which modulates the navigation message. The value of this term depends on the used modulation technique changing according to the type of signal.
- $f_{RF,i}$ is the nominal carrier frequency.
- f_D^i is the Doppler frequency of the i^{th} received signal.
- $\varphi_i(t)$ represents the phase variations of the i^{th} received signal.

GNSS signals are transmitted on the L band, which is a part of the frequency spectrum ranging from 1 up to 2 GHz. This frequency band has been allocated by the International Telecommunication Union (ITU) to the Radio Navigation Satellites Services (RNSSs) to protect this band from being interfered by other communication

systems. Furthermore, the L band allows to achieve little atmospheric attenuation even if, as detailed in Chapter 3, it is still subject to ionospheric delays and scintillation (Bhaskar, 2015). Figure 2-3 shows the present and planned frequency allocation for GPS, Galileo and GLONASS. In particular, it can be seen that each GPS signal transmits data centred on three frequencies, namely L1=1572.42 MHz, L2=1226.70 MHz and L3=1176.45 MHz. L1 and L2 share the band with the Aviation Radio Navigation Service (ARNS) which offers the advantage of enjoying a strong protection solution against interference and, consequently, is particularly indicated to provide safety related services. The algorithms proposed by this thesis will be assessed by exploiting GPS L1 C/A signals, although, as explained later, the same approach could also be adopted for other frequencies. Among them the GPS L1 C/A is the only signal including the civilian Coarse Acquisition code. The RF carrier centred at the frequency determined by the type of signal is modulated with the navigation message and the PRN code, which, as previously underlined, includes the information required by the receiver to compute the satellite position.

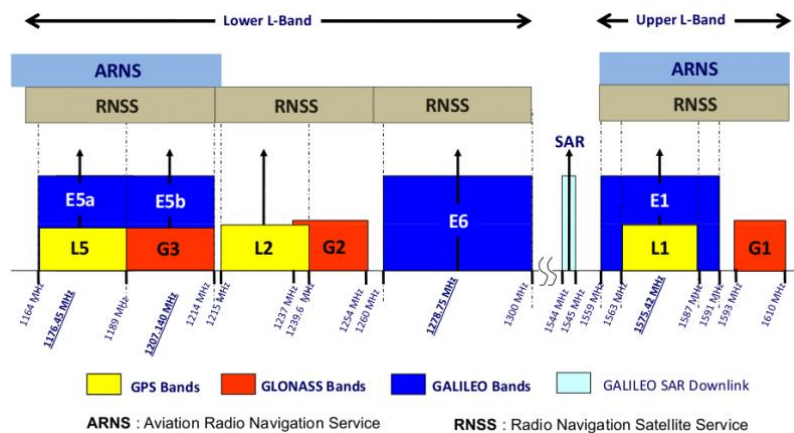


Figure 2-3: GNSS frequency plan allocation.

Navigation data bit

The signal carrying the navigation data can be expressed as:

$$d_i(t) = \sum_{n=-\infty}^{\infty} d_i[n] g_{T_b}(t - nT_b) \quad (2.3)$$

Where $d_i[n]$ is the sequence of data symbols, T_b is the data bit duration and $g_{T_b}(t)$ is a rectangular pulse defined as:

$$g_{T_b}(t) = \begin{cases} 1 & 0 < t < T_b \\ 0 & elsewhere \end{cases} \quad (2.4)$$

In the case of GPS L1 C/A signals T_b is equal to 20 ms. The navigation message includes all the information required by the receiver to compute the position of all satellites in view and the time of signal transmission.

Spreading Code

The spreading code is a periodical binary Pseudo Random Noise (PRN) sequence, which is multiplied by the input signal to increase the bandwidth of the original signal. Each satellite has its own specific code, which can be expressed as:

$$c_i(t) = \sum_{n=-\infty}^{\infty} c_i[n] g_{T_{chip}}(t - nT_{chip}) \quad (2.5)$$

Where $c_i[n]$ is the periodic repetition of the code sequence, T_{chip} is the duration of a spreading code pulse, termed as chip and $g_{T_{chip}}(t)$ is the rectangular pulse. In the case of GPS L1 C/A signals the code length is of 1023 chips and T_{chip} is equal to 1 ms. GPS signals exploit the so-called Gold codes produced by modulo 2 addition of two maximal length sequences (m-sequences). These types of codes are selected for their good correlation properties. Indeed, they show high auto-correlation peaks and low cross-correlation peaks (Pany, 2010). This property is fundamental for the demodulation process. In fact when the correlation is performed between the input signal and a synchronized local replica of the code the correlation output should be large indicating an autocorrelation peak. When the input signal is correlated with a different PRN code the correlation output should be lower. Using orthogonal codes a cross correlation equal to zero can be obtained. The Gold codes, exploited in the transmission of GPS signals, are quasi-orthogonal and consequently they produce cross correlation values very close to zero (Misra and Enge, 2011). An example of auto correlation for a PRN sequence used in GPS signals is shown on the left side of Figure 2-4 where the sharp peak is clearly visible, while a sample case of cross correlation function is reported on the right side of Figure 2-4.

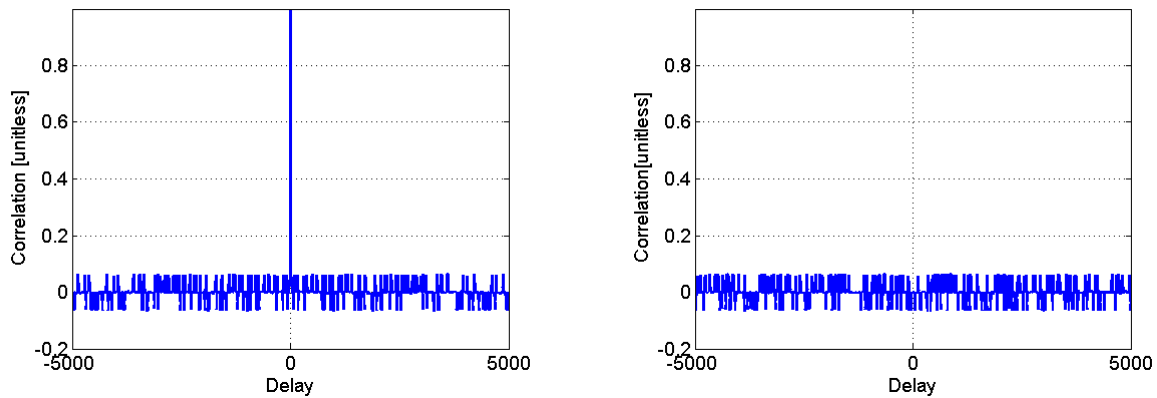


Figure 2-4: (Left) Auto correlation example for a Gold Code employed by the GPS C/A code; (Right) Cross correlation example for a Gold Code employed by the GPS C/A code.

2.5 General structure of a GNSS receiver

When a GNSS signal arrives at the antenna it is buried into the noise so the signal needs to be processed in order to estimate the synchronization parameters necessary to estimate the PVT solution. This is performed by the different stages of the receiver, which are represented in Figure 2-5. After being received by the antenna, the front end filters, amplifies and down converts the signal to an intermediate frequency (IF) and then digitizes this IF signal so as to allow the signal processing to be performed by the following receiver stages. Then, the acquisition block detects all satellites in view and provides a rough estimate of code delay and Doppler frequency. These estimates are then refined by the tracking stage which can, optionally, track the carrier phase offset. In standard receivers, the tracking stage is composed of two interleaved tracking blocks, namely the code and the carrier tracking. Code tracking is performed through a DLL while the carrier can be tracked by using a FLL or/and a PLL. Specifically, if a non-coherent tracking architecture is exploited, only the code delay and Doppler frequency are tracked and a FLL is used for

the carrier tracking. When a coherent architecture is adopted, as in this research work, code delay, Doppler frequency and phase offset are tracked and the signal carrier is tracked by using a PLL (Misra and Enge, 2011). Moreover the tracking of these parameters should be performed over the time to follow the signal variations. When the signal code and carrier are tracked, the navigation data bits can be extracted and the pseudoranges can be estimated. Therefore the measurements provided by the tracking stage (i.e. Doppler frequency, carrier phase, pseudoranges) are then input to the PVT block (Kaplan and Hegarty, 2006) which estimates the user's position, velocity and time. A more detailed description of these different blocks is reported in the following paragraphs.

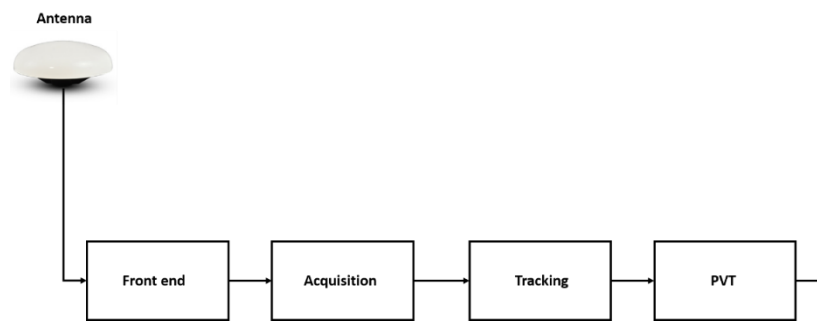


Figure 2-5: General architecture of a standard GNSS receiver for a single channel.

2.5.1 Antenna

The first stage of a GNSS receiver is the Right-Hand Circularly Polarized (RHCP) antenna, the purpose of which is to receive the GNSS signals broadcasted in the L band. A GNSS antenna, when properly designed, must maximize its gain in the direction of the satellites in view while minimizing the gain along the direction of reflected (or so-called multipath) signals, such as the ones arriving from low elevations. An essential feature, which determines the antenna performance, is its phase centre, which is the reference for the

receiver to estimate its position. In fact the position estimated by the receiver is the position of the antenna electrical phase centre, which is the point of the antenna where the signal is received. Depending on the design of the GNSS antenna, the electrical phase centre may change according to the arrival directions of the signal and it must be accurately determined to achieve accurate solutions (Kaplan and Hegarty, 2006).

2.5.2 Front-end

The front end has the purpose of amplifying, filtering and down converting the incoming signal to the IF that can then be converted from the analogue to the digital domain through an Analog Digital Converter (ADC), which is the last front end block. The structure of a generic GNSS front end is presented in Figure 2-6 (Misra and Enge, 2011). The first stage is a Band Pass (BP) filter which rejects all the out of band interference, while maintaining the GNSS waveform. Then, since the signal arriving at the receiver antennas is extremely weak, it needs to be amplified in order to be processed by the later blocks. For this purpose a Low Noise Amplifier (LNA) is employed. The LNA should be characterized by the highest gain and the minimum noise so as to minimize the chain overall noise figure according to the Friis formula (Haykin, 2001). After amplification the signal can be converted to an IF enabling the next stages of the signal processing. Down conversion is performed through a mixer. The latter multiplies the signal by a locally generated carrier signal at frequency f_{LO} produced by a local oscillator to obtain a resulting signal centred at IF with $f_{IF} = f_{RF} - f_{LO}$.

The signal from one satellite converted to IF at the output of the front end can be expressed as (Misra and Enge, 2011):

$$r_{IF}^i(t) = \sqrt{2P_i(t)}d_i(t - \tau_i)c_i(t - \tau_i)\cos\left(2\pi(f_{IF} + f_D^i)t + \varphi_i(t)\right) + n_{IF}(t) \quad (2.6)$$

It is worth mentioning that thanks to the quasi orthogonality of the spreading codes, the receiver can process each single component in an independent way. Therefore, in the following of this thesis the signal will be expressed as a single component. The signal in (2.6) is further filtered by an IF Band Pass Filter (BPF) in order to remove the high frequency components, also defined as image frequency components, introduced by the multiplication operation required by the down conversion process.

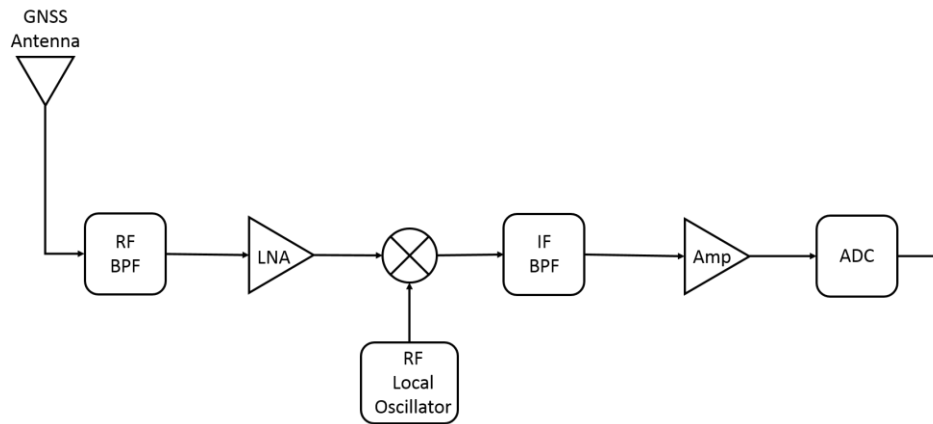


Figure 2-6: General structure of a GNSS front-end.

Finally, the analogue signal is input to the ADC block, which samples and quantizes the signal to obtain a digital version of the signal. The digital signal at the front-end output can be expressed as:

$$r_{IF}^i(nT_s) = \sqrt{2P_i(nT_s)}d_i(nT_s - \tau_i)c_i(nT_s - \tau_i)\cos\left(2\pi(f_{IF} + f_D^i)nT_s + \varphi_i(nT_s)\right) + n_{IF}(nT_s) \quad (2.7)$$

Where

- f_{IF} is the new centre frequency
- $T_s = \frac{1}{f_s}$ is the sampling interval exploited to obtain the digital sequence and f_s is the sampling frequency
- f_D^i is the Doppler frequency
- φ_i is the carrier frequency
- n_{IF} is the noise affecting the signal. This term also includes the receiver thermal noise produced by the various front-end blocks. Assuming that no interference or spurious signals occur this term is well modelled as a white Gaussian noise with a double side PSD equal to $N_0/2$ [W/dBHz].

The above digital signal can also be expressed in its discrete form as

$$r_{IF}^i[k] = \sqrt{2P_i[k]}d_i[k - \tau_i]c_i[k - \tau_i]\cos\left(2\pi(f_{IF} + f_{D,i})[k] + \varphi_i[k]\right) + n_i[k] \quad (2.8)$$

Where $k = nT_s$.

2.5.3 Baseband Signal Processing

The baseband signal processing block has the purpose to demodulate the GNSS signals provided by the front-end by correlating them with locally generated copies of the

spreading code and carrier. As mentioned above, this task is performed through two receiver stages, namely the signal acquisition and tracking. The acquisition detects the signal presence and provides rough estimates of code delay and Doppler frequency. Then the tracking refines these estimates, and optionally provides the carrier phase offset. Once these parameters are estimated, the signal carrier and code can be wiped off in order to extract the navigation bits and construct the measurements necessary to the PVT block. More details on the acquisition and tracking stages are provided in the following paragraphs.

2.5.3.1 Acquisition

The acquisition block has the objective to detect the signal presence and provide a list of all satellites in view through the so-defined sky search process. Then, for each detected satellite, the related code delay τ and Doppler shift f_D are roughly estimated. Considering each single satellite the acquisition can be seen as a 2 dimensional search process. The two-dimensional search space, represented in Figure 2-7, is produced by correlating the input signal with different versions of the locally generated code. These different code versions are obtained by translating the local signal replica in delay and in frequency to recreate a grid covering all the possible delay and Doppler shift ranges. As shown in Figure 2-8, the correlation between the input signal and its local replica shows a sharp peak if the code delay and Doppler frequency of the input signal match with the ones of the local replica. Consequently, when the amplitude of this peak is above a predefined threshold

the presence of a satellite can be flagged. Details on different rules to define this threshold are reported by Borio (2008).

Therefore, when a satellite is detected the code delay and Doppler shift of the correspondent signal replica are used as rough estimates, which are then provided to the tracking block to be refined. Different acquisition algorithms can be implemented according to the approach used for the search space. The accuracy of the search process is inversely dependent on the size of the time and frequency steps used to build the above search space grid. The smaller this step is the higher is the accuracy, although this will be at the cost of an increased computational load. The frequency range for the search process is strongly determined by the expected signal dynamics and the quality of the receiver clock. For instance, for classical terrestrial receivers, the frequency range is usually up to ± 10 kHz while for space based receivers a wider frequency range should be adopted, up to ± 50 kHz (Blunt, 2007) due to the increased speed generally being used in space vehicles. Furthermore, there are different techniques to perform this space search process. The most common approach is based on the computation of the Fast Fourier Transform (FFT), which allows searching in parallel Doppler frequency and code delay so as to make the process faster. More details on acquisition algorithms can be found in (Misra and Enge, 2011).

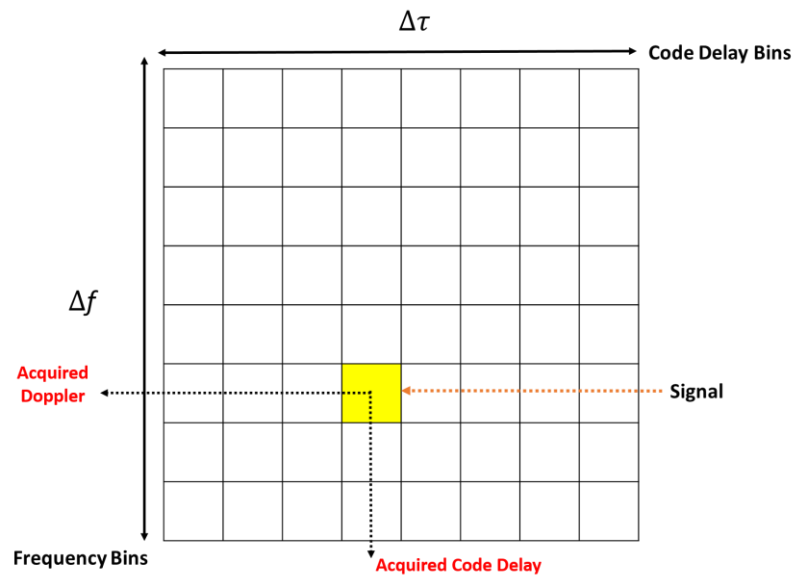


Figure 2-7: Representation of a two-dimensional acquisition search space. The filled square indicates the acquired Doppler and code delay cell.

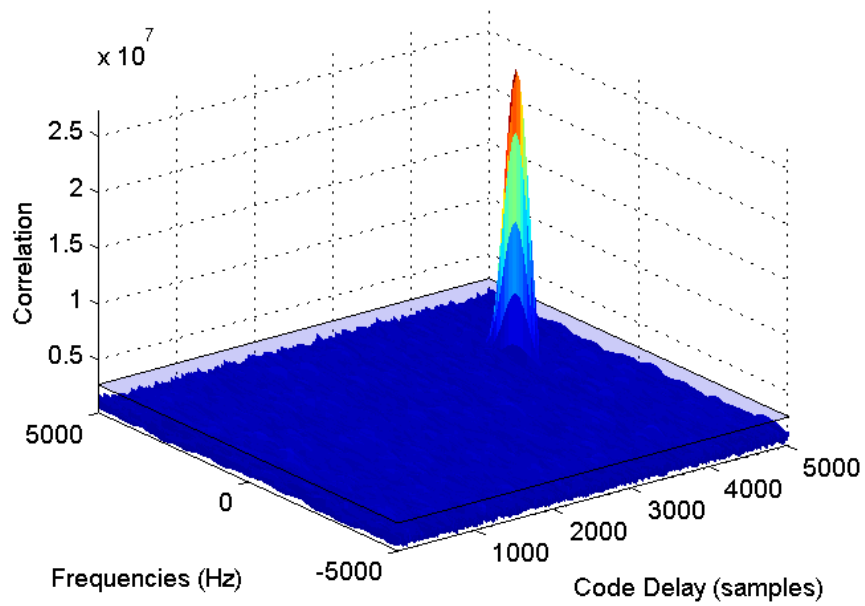


Figure 2-8: Sample case of correlation output for a satellite in view.

2.5.3.2 Tracking

Once the acquisition stage provides the list of all satellites in view and the related rough estimates of the code delay and the Doppler frequency the acquired information is then input to the tracking block. As previously mentioned, the tracking stage refines these estimates and, in addition, can track the carrier phase offset. Moreover, this receiver stage is in charge of tracking these synchronization parameters over time and demodulate the navigation message so as to generate the measurements necessary to compute the PVT solution. In a traditional scalar architecture the tracking is performed for each channel separately. However, also tracking vector options, which track more satellite links jointly can be adopted. These types of architecture will be detailed in Section 5.2. Conventionally, in order to track the signal code and carrier, respectively, a Delay Locked Loop (DLL) and a Frequency Locked Loop (FLL) or/and a Phase Locked Loop (PLL) are employed. It is worth underlining that apart from tracking schemes based on closed loop structures, also open loop architectures, described in Section 5.2, could be used. DLL, FLL and PLL are characterized by the same general structure, which is shown in Figure 2-9.

In the carrier tracking loop case, the Numerically Controlled Oscillator (NCO) drives the generation of a local replica of the carrier signal while in the code tracking case the NCO drives the speed of the replica code generator. Then the signal replica is correlated with the input signals and integrated over a certain temporal window by the Integration and Dump (I&D) block. The integration in a discrete domain is equivalent to a sum over N samples, as indicated in Figure 2-9. The duration of the temporal window used for the integration is indicated as time of integration. The integration has the purpose of reducing the noise, being the integration operation equivalent to a low pass filter. Then the *I&D*

output is fed to the non-linear discriminator. This stage provides an estimate of the error in the parameter to be tracked. Then the loop filter has the objectives of minimizing the noise of the discriminator output and following the signal dynamics. Finally, the NCO adjusts the replica of the code or carrier according to the filter output. In Figure 2-9 $k[n]$ indicates the parameter to be tracked, $\tilde{k}[n]$ is the estimate of the above parameter provided by the loop and $\varepsilon[n]$ is the estimation error. The difference between DLL, FLL and PLL is determined by the parameter to be tracked, which will result in a different type of non-linear discriminator. More details about these tracking architectures will be provided in the following sections.

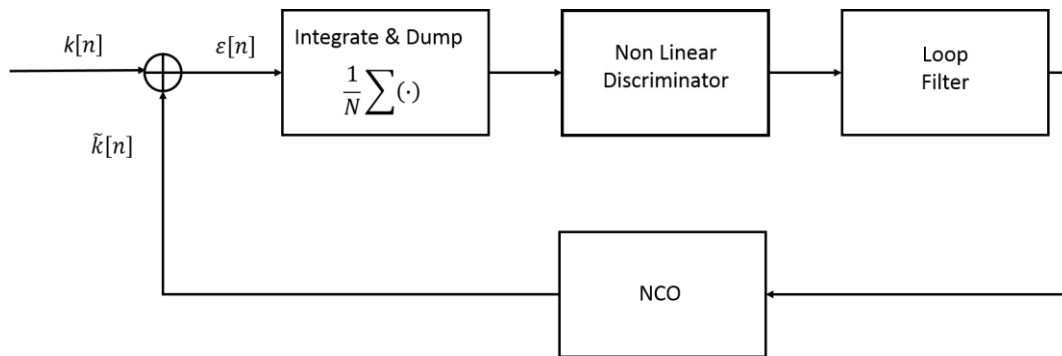


Figure 2-9: General structure of a tracking loop.

2.5.3.2.1 Delay Locked Loop

The DLL, whose general scheme is shown in Figure 2-10, is exploited to track the code delay by maximizing the correlation between the input signal and the locally produced code (Misra and Enge, 2011). This goal is achieved by multiplying the input signal with the In-phase (I) and in Quadrature (Q) versions of the locally produced carrier provided by the

carrier tracking in order to wipe off the code phase. Then the resulting signal is multiplied by three replicas of the code phase with different delays. They are defined as:

- **Early code (E)**, This is produced by advancing the code with respect to the present time instant by an amount equal to $d/2$, with d equal to the space chip.
- **Prompt code (P)**, which is generated without any delay.
- **Late code (L)**, which is produced by delaying the code with respect to the present time instant by an amount equal to $d/2$.

The output of this block is fed to the aforementioned I&D block, which provides the result of the correlation between the code of the input signal and its locally produced replicas. The output of the I&D block is provided to the code discriminator, which extracts the code estimation error. Discussions about different types of DLL discriminators can be found in Ward (2006). Then the noise affecting the output of the discriminator is reduced through the loop filter. In order to produce a local replica of the code synchronized with the one of the input signal, the output of the early and late correlators are compared so as to make the power of these two channels equal, as shown in Figure 2-11.

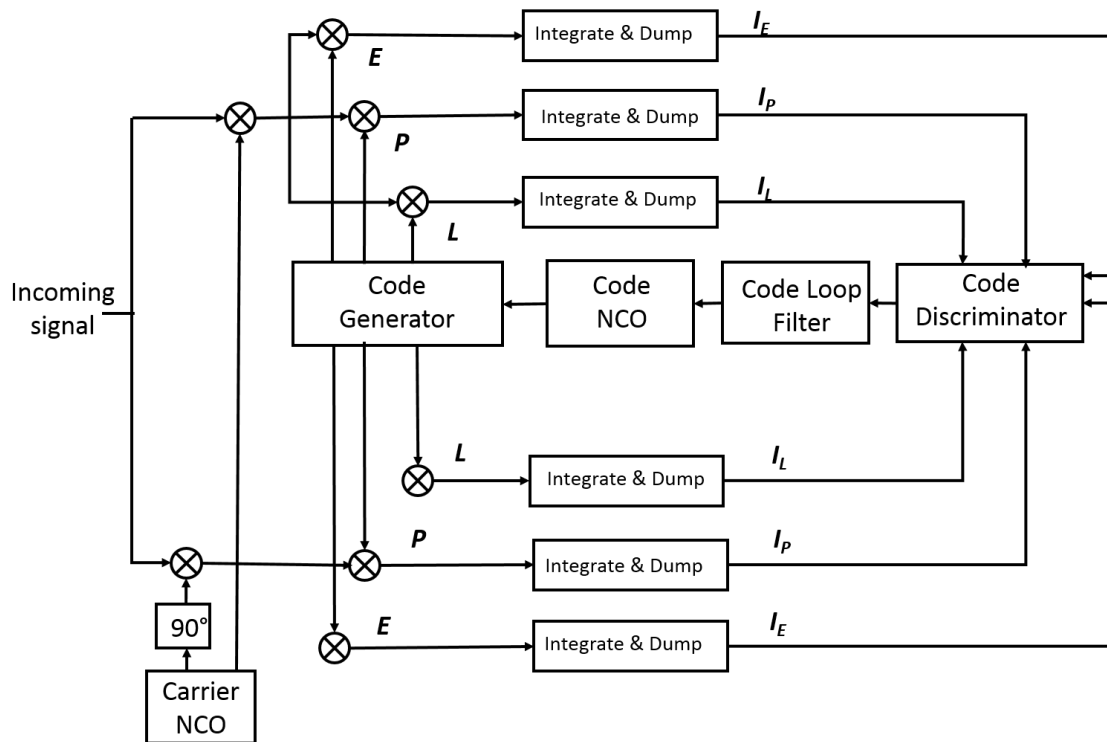


Figure 2-10: Diagram of a DLL (modified from Borre, 2007).

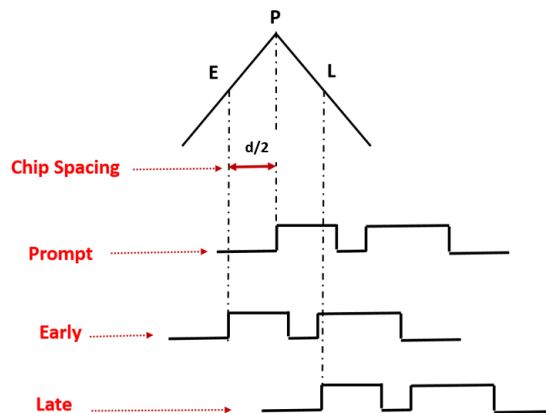


Figure 2-11: Illustration of the code delay tracking with prompt, early and late arms.

2.5.3.2.2 Carrier tracking loop

The tracking of the carrier signal can be performed by exploiting a Frequency Locked Loop (FLL) or/and a Phase Locked Loop (PLL). The former tracks only the carrier frequency while the latter estimates both phase and carrier frequency. In particular, the PLL, whose structure is illustrated in Figure 2-12, detects the phase discrepancy between the input signal and the local replica.

Since the initial phase of the input signal is not known, the NCO produces two versions of the local carrier shifted of 90° so as to obtain the I and Q components. The input signal is then multiplied by the local replica of the code to wipe off the code and subsequently integrated over the integration time to reduce the noise. The output of the discriminator block is fed to the loop filter to extract the phase information. Details about different types of PLL discriminators can be found in (Kaplan and Hegarty, 2006). The discriminator provides an estimate of the phase error that is then input to a loop filter to improve the quality of this measurement. Then the loop filter output is used to drive the NCO and consequently to adjust the produced frequency. A standard PLL is sensitive to data bit transitions, namely changes in the phase of signal equal to 180° . To overcome this issue, a specific class of discriminators, indicated as Costas discriminators, can be used in order to ensure insensitivity to data bit transitions (Kaplan and Hegarty, 2006).

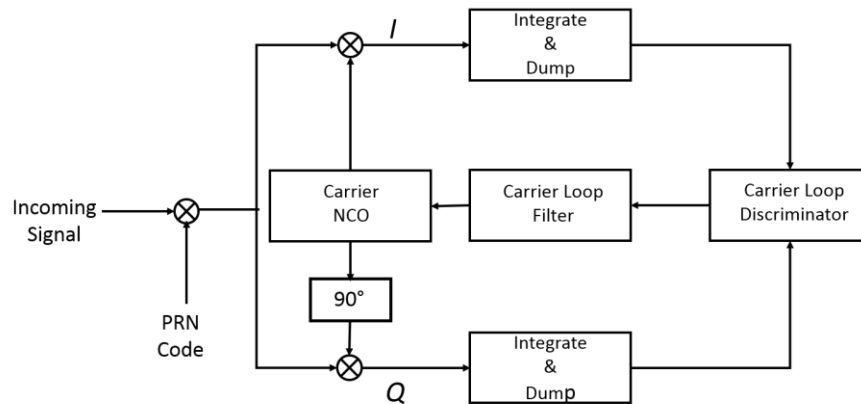


Figure 2-12: Diagram of a PLL (modified from Borre, 2007).

The FLL has a structure very similar to the PLL. However, this architecture only provides an estimate of the frequency, without giving information about the carrier phase. To extract the error in frequency a specific FLL discriminator should be exploited. Different types of FLL discriminators are reviewed by Kaplan and Hegarty (2006). FLL discriminators exploit the values of the phase at the current and previous instant in order to compute the absolute phase difference between two epochs, while PLL discriminators only require the phase at the present epoch. Moreover, the FLL architecture also includes an integrator in order to integrate the loop filter output and estimate the phase rate. In a standard GNSS receiver tracking loop, DLL and PLL/FLL work in an interleaved way as shown in Figure 2-13, where the complete structure of a GNSS receiver tracking loop is presented. To summarize, the DLL produces a local replica of the code delay, which is used to wipe off the code from the signal. In this way a pure carrier can be input to the PLL/FLL. At the same time the PLL/FLL is used to produce a local replica of the carrier so as to wipe off the carrier from the input signal allowing the DLL to track the code delay.

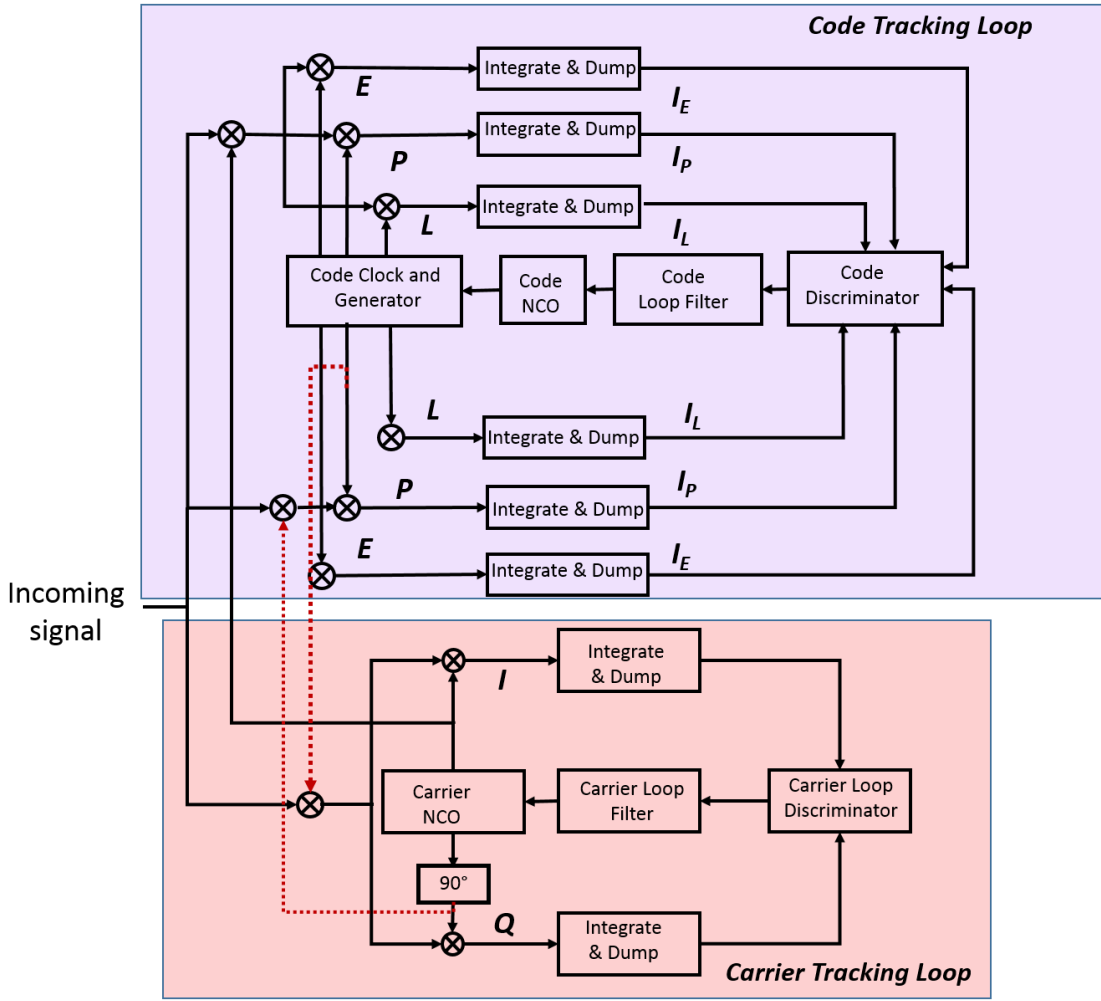


Figure 2-13: GNSS receiver tracking loop (modified from Borre, 2007).

The estimated code delay can be used to construct the pseudorange measurements which are defined as follows

$$\rho_i = r_i + c(t_{Rx} - t_{Tx,i}) + \varepsilon_i \quad (2.9)$$

Where r_k is the geometric range $\|r_{Rx} - r_{Tx,i}\|$ between the receiver and the satellite i , c is the speed of the light, t_{Rx} is the receiver time, $t_{Tx,i}$ is the signal transmission time and ε_i is

the error term produced by different error sources which will be reviewed in Section 2.6. In order to estimate the pseudorange it is necessary to recover the transmission time from the navigation message of the received signal. For this purpose first of all the bit boundaries has to be identified through a data bit synchronization process. Then the demodulation of data bits can be performed. Different techniques to perform data bit synchronization and decoding can be found in (Spilker, 1997). Once the bits have been extracted the signal transmission time, embedded in the navigation message, can be recovered. The GPS L1 C/A navigation message includes a time mark, termed as Z count, which is transmitted every 6 seconds with its value incremented each 1.5 seconds. To estimate the satellite clock time corresponding to the current epoch the Z count should be summed to the number of navigation data bits (20 ms), the number of C/A code periods (1 ms), the number of whole C/A code chips and the fraction of the current code chip. More details on different techniques for the pseudorange computation are discussed by Pini et al. (2012).

2.5.4 PVT

The measured pseudorange between the receiver and satellite i , defined by equation (2.9), can be rewritten by expanding the geometric term as:

$$\rho_i = \sqrt{(x_s^i - x_u)^2 + (y_s^i - y_u)^2 + (z_s^i - z_u)^2} + c(t_{Rx} - t_{Tx,i}) + \varepsilon_i \quad (2.10)$$

Where $P_u = [x_u \ y_u \ z_u]$ defines the receiver position and $P_s^i = [x_s^i \ y_s^i \ z_s^i]$ indicates the position of the i^{th} satellite. To compute the satellite's position the receiver exploits the information retrieved by the navigation message, such as ephemeris and time of transmission. The pseudorange expression includes four unknown, namely the three position coordinates and the clock bias. Consequently, a system of, at least, four equations is necessary to estimate the user's position. Since the relationship between the measured pseudorange and the user's coordinates is non-linear, the standard approach to solve this system of equations is to use iterative techniques, such as least square or KF based estimators. Indeed these techniques enable the linearization of the set of equations around an initial coarse guess of the user's position and clock bias. However this research work does not deal with the receiver navigation block, but focuses only on the receiver tracking stage. Therefore, for an extensive review of this subject interested readers can refer to Misra and Enge (2011), Kaplan and Hegarty (2006).

2.6 GNSS Error Sources

In a general form, the pseudo-range accuracy is affected by three types of error sources, which can be classified as:

- Satellite based errors
- Propagation errors
- Receiver errors

A typical scenario with the main error sources affecting GNSS is illustrated in Figure 2-14.

A short overview of these error sources is given in the following sections while for a more

detailed description the reader can refer to (e.g. Kaplan and Hegarty 2006, Misra and Enge 2011).

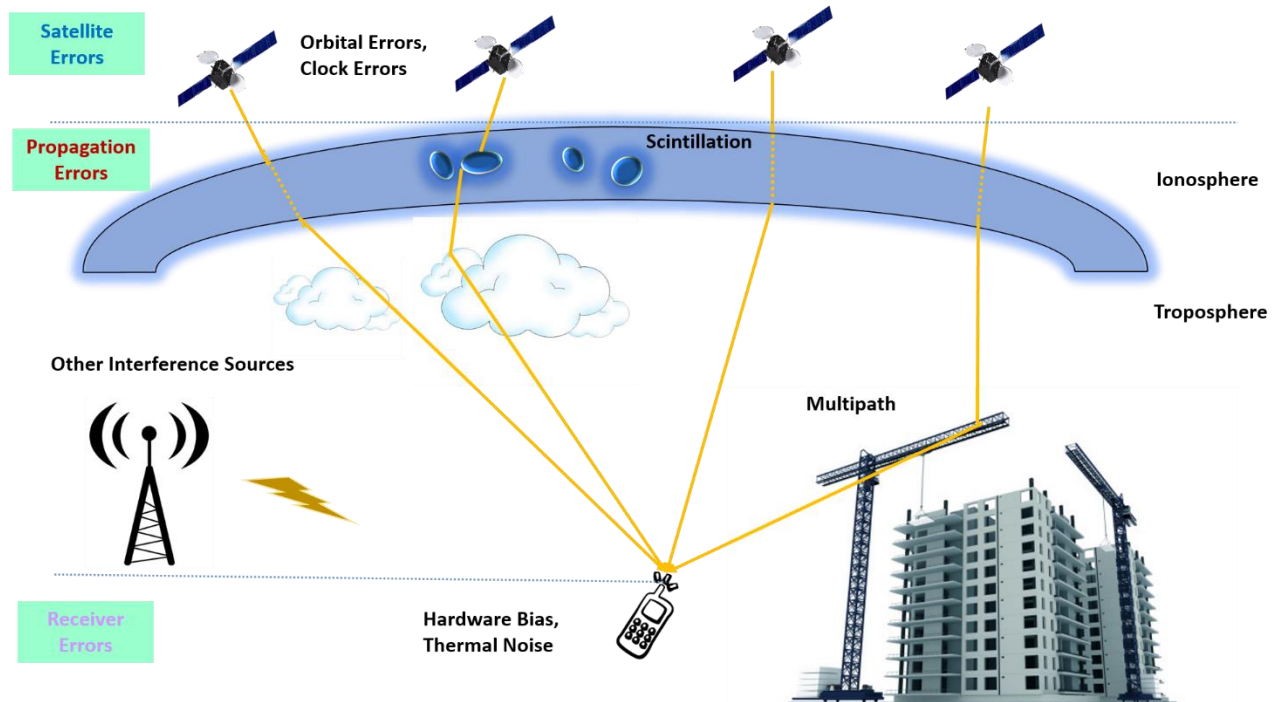


Figure 2-14: Illustration of a typical scenario including the main GNSS error sources.

2.6.1 Satellite based errors

Satellite based errors refer mainly to satellite orbit and clock errors. Satellite orbit errors are errors in the estimated satellite position from the orbit parameters included in the ephemeris. Satellite clock errors are given by the discrepancy between the time provided by the atomic clocks on board satellites and an established reference time system. In fact even if the clocks on board satellites are very accurate and expensive atomic clocks, still a difference with the reference time can occur. The GCS estimates clock correction parameters that are then transmitted to the satellites. These parameters, similarly to those

for the ephemeris, are then included in the navigation data broadcasted with the signals from the satellites. Both the satellite orbit and clock errors can be largely removed by applying differential techniques between receivers.

2.6.2 Signal propagation errors

GNSS signals broadcasted by satellites must travel through a physical transmission medium, defined as a communication channel, before being received by the receiver antenna. The characteristics of this communication channel are time and space variable and, consequently, affect the signal propagation in different ways. The first potential source of propagation error encountered by GNSS signals in their path is represented by the atmosphere, and especially by two of its layers, namely the troposphere and the ionosphere, which affect GNSS signals in different ways. The troposphere, the neutral and lower layer of the atmosphere (up to 20km in altitude), produces a delay in the signal propagation which is a function of its refractive index, the temperature, the humidity and the pressure. The modelling of these tropospheric errors allows mitigating their effects in a largely effective way. The ionosphere, the layer of the atmosphere extending from about 60 km up to 1000 km above the earth surface, is a dispersive medium since its refractive index changes according to the frequency of the propagating signal frequency (Ghaafori, 2012).

The variation of the refractive index in the ionosphere affects the propagation speed of GNSS signals, producing as a result a code delay and a phase advance of the signals. These ionospheric effects are predictable and in fact they can be mitigated by applying specific ionospheric correction models when single frequency measurements are

exploited. Moreover, considering the dispersive nature of the ionosphere, in the case of dual frequency receivers, the above effects can be removed by combining GNSS measurements at different frequencies.

As previously mentioned, the ionosphere can also affect the signal propagation in a random way through scintillation events, producing fast and unpredictable variations in the signal amplitude and phase. The mitigation of scintillation events through a proper design of the tracking stage is the main objective of this thesis and the proposed novel algorithms robust under scintillation will be detailed in Chapter 6. Furthermore, the effects of the ionosphere on GNSS signals will be reviewed in more depth in the following chapter, entirely dedicated to this topic.

Another consistent source of error is represented by signal multipath. Multipath refers to the arrival of the signal at the receiver antenna via multiple paths. As results the signal can be strongly attenuated or its phase can be subjected to rapid variations. This error source represents a major challenge, especially in urban canyons and in indoor environments where the presence of obstacles affects the GNSS signal propagation. The level of the multipath range measurement error is determined by the strength of the reflected signal and the delay between reflected and direct signals (Kaplan and Hegarty, 2006). The most intuitive way to mitigate multipath is to place the antenna in a location far from reflectors. When this is not possible a proper antenna design or specific receiver tracking architectures can help to mitigate multipath effects.

Finally, different types of interference can affect the performance of a GNSS receiver. RF interference, namely any undesired signal receipted by the receiver, (Kaplan

and Hegarty, 2006) can be classified, on the basis of its source, as intentional, including spoofing and jamming, and unintentional. Unintentional interference could be produced, for instance, by any communication system transmitting at carrier frequencies close to the ones of the signal of interest (Dovis, 2015). Moreover RF interference can be distinguished as Narrow Band (NB) and Wide Band (WB) according to the dimension of the bandwidth of the interfering signal with respect to the one of the desired GNSS signal. NB interference can produce strong harmonics and intermodulation products (Bhaskar, 2015) which affect the GNSS signal detection and tracking. WB interference can compromise the GNSS signal detection by increasing the GNSS noise floor. According to the type of interference, different approaches can be used to detect and mitigate the interfering signal at antenna, pre-correlation or post-correlation levels. More details about RF interference and related mitigation strategies can be found in (Dovis, 2015).

2.6.3 Receiver errors

These types of errors are generated inside the receiver and they are mainly caused by hardware biases and thermal noise. The level of this class of errors strongly depends on the technology used inside the receiver. In the case of high quality receivers these errors can be neglected for the carrier phase, while they can produce some decimetres of errors for the code phase. Hardware biases refer to the delay produced by receiver components, such as antennas, cables, filters, LNAs and mixers. The latter type of errors affect in equal measure all signals received by the receiver antenna. Consequently, they can be modelled as a bias factor in the navigation solution block.

2.7 Summary

This chapter provided a general background to GNSS. In particular, since the algorithms of this thesis aim to mitigate ionospheric scintillation effects through a novel design of the receiver's tracking stage, particular emphasis has been given to the description of the structure of a GNSS receiver, and especially to the tracking stage, which is the receiver's link that is more vulnerable to scintillation effects. Finally, the main error sources affecting GNSS accuracy are described. The next chapter will be completely dedicated to the ionosphere which is one of the main contributors to the total GNSS error budget. In particular, the ionospheric scintillation and its impact on the tracking loop of a GNSS receiver will be described.

Chapter Three: Ionospheric Morphology and Scintillation Events

This chapter describes the structure of the ionosphere, detailing how this atmosphere's layer affects the propagation of GNSS signals. The chapter will centre especially on the scintillation phenomenon, namely the random signal and phase variations produced by small-scale structures, which may occur inside the ionosphere. Scintillation characteristics strongly depend on different temporal and spatial factors, which determine the make-up of the ionosphere. They include the solar and geomagnetic activity, season, local time, signal frequency, and geographic location. This chapter describes how the above factors determine scintillation, focusing on the different ionospheric morphology of polar and equatorial regions, which lead to different scintillation characteristics. Finally, different approaches to model scintillation will be briefly reviewed.

3.1 The ionosphere

As introduced in Chapter 2, the ionosphere is the layer of the upper atmosphere, ranging approximately from about 60 km up to 1000 km over the Earth. In contrast to the lower atmosphere layers, which are neutral, the ionosphere is an ionised medium, because of the presence of free electrons produced through the ionisation of the neutral gases. This is due mainly to the Sun's high-energy radiation, but also to the collision with cosmic particles. Since the charged particles are fewer than the neutral particles, the ionosphere is considered as a weakly ionised plasma (Tascione, 1988).

However, these charged particles determine the electrical properties of the ionosphere and affect the propagation of trans-ionospheric radio wave signals. The ionisation process

is mainly produced by the high-energy solar radiations, in the form of Ultra Violet (UV) and X rays, which, by penetrating in the atmosphere, heat the gas molecules and induce their dissociation with the consequent production of negatively charged free electrons and positively charged ions (Davies,1990;Tascione, 1988;Kelley, 1989).

A recombination process can also occur when, after random collisions, these free electrons are captured by the free ions. Consequently, the balance of ionisation and recombination determines the level of the final ionospheric ionisation. The rate of these two processes is mainly driven by the Sun's incidence and the atmosphere's composition. The solar radiation increases while the atmosphere density decreases with the altitude. Moreover, the absorption characteristics of the ionosphere change with the height. All these factors lead to a vertical stratification of the ionosphere.

Typical day and night time curves for the electron density in the case of a mid-latitude location is reported in Figure 3-1. The two continuous curves represent the electron density profiles during the day and at night time under quiet conditions, while the two dashed lines represent the electron density curves during the night and in daytime hours, but under turbulent conditions.

It can be observed that the bottom side of the ionosphere is characterised by the layers indicated as *D*, *E*, and *F*. *F* marks the beginning of the upside of the ionosphere, which is characterised by a monotonic decrease of the electron density, due to a minor density of the ionosphere, and consequently, to a minor number of electrons, which can be ionised at those heights (Davies, 1990; Barcley, 2003).

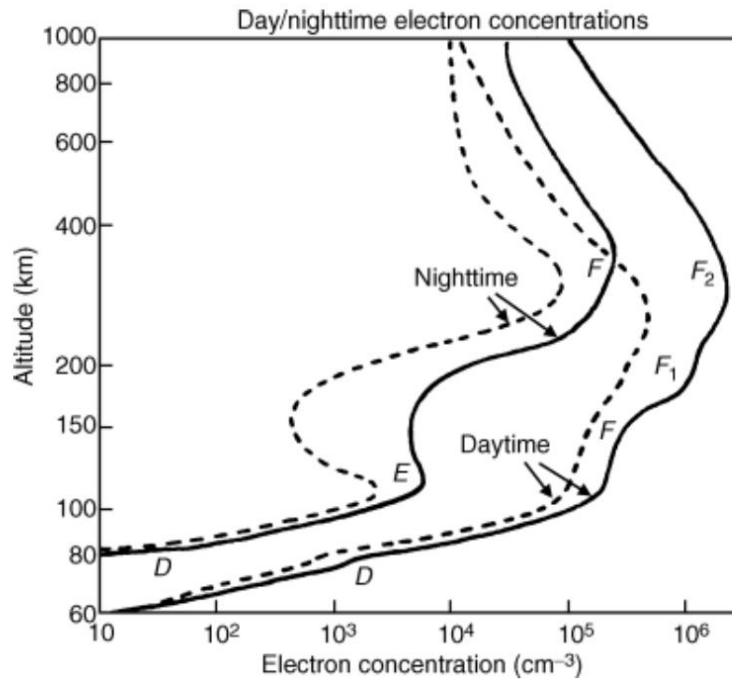


Figure 3-1: Sample case of vertical layer of the ionosphere at mid latitudes during night and day under quiet (continuous lines) and turbulent (dashed lines) conditions (extracted from Barclay, 2003).

The **D layer** is the lowest part of the ionosphere extending from about 50 km up to 90 km. The lowest part of the D layer is usually indicated as the C layer to distinguish this part, only produced by the cosmic radiations, from its upper part, which is produced by both cosmic and solar radiations. At night time, since only the recombination process occurs, this layer is strongly reduced, and it can even disappear.

The **E layer** goes from around 90 km to 140 km above the Earth, and is mainly produced by the action of the solar UV radiation. It appears during the day, while it is subjected to consistent reduction during the night.

Finally, the **F layer** is the one with the major concentration of electrons and it can be divided into the two following sublayers.

- The **F1 sublayer** ranges approximately from 160 km to 200 km and, under quiet ionospheric conditions, it usually appears only during the summer at daytime.
- The **F2 sublayer** stretches from 210 km to 1000 km. This is the layer of the ionosphere with the maximum concentration of electrons and that, consequently, has the bigger influence on the propagation of GNSS signals. Moreover, it is characterised by the highest variability, both in space and in time. While at night the F1 disappears, the F2 still persists.

Due to the variation of sun radiation intensity and magnetic activity, the ionospheric ionisation level is highly variable both in space and in time (Barclay, 2003). Indeed, by comparing the dashed and continuous profiles in Figure 3-1, we can see that during the daytime the electron concentration is bigger if compared with that at night time. However, since the Sun's incidence changes also with the season and according to the solar cycles, this also affects the ionosphere's morphology. The main factors affecting the electron density variations are briefly described in the following sections.

3.1.1 Geomagnetic latitude variations

The electric currents flowing at the core of the Earth produce a geomagnetic field, which can be approximated (Tascione, 1988) as a sphere magnetised along the direction of a dipole axis in a uniform way. As shown in Figure 3-2, the dipole axis stretches from the North (or austral) to the South (or boreal) poles and it is inclined at about 11° with respect to the rotation axis of the Earth. By intersecting the plane perpendicular to the

dipole axis, and traversing the centre of the Earth with the Earth surface, the geomagnetic equator, which is different from the geographical equator and is characterized by a horizontal magnetic field, can be obtained.

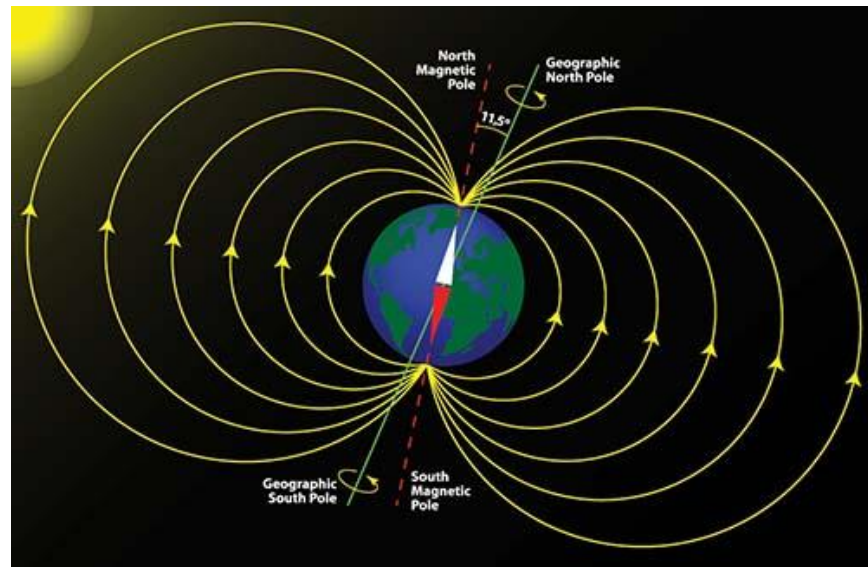


Figure 3-2: Earth magnetic dipole representation.

Three main geomagnetic regions can be identified. They are the equatorial, mid-latitude, and high latitude regions, which are represented in Figure 3-3 (Odijk, 2002).

The Earth's magnetic field strongly affects the morphology of the ionosphere and, consequently, these three geomagnetic areas also define three different ionospheric regions, with different ionospheric properties described as follows.

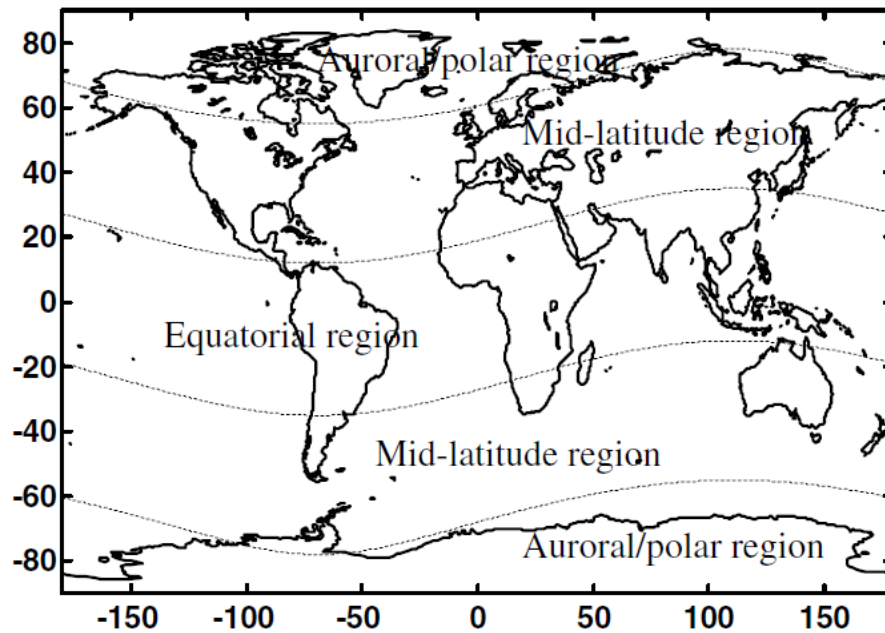


Figure 3-3: Earth geomagnetic regions (extracted from Odjik 2002).

3.1.1.1 Equatorial region

The equatorial region, also indicated as low latitudes, stretches from about 15° - 20° degrees above and below the geomagnetic equator. This region includes the so- termed equatorial anomalies, or Appleton anomalies, which are located at about 10° - 20° degrees above and below the equator, and are characterised by the highest values of electron density on the Earth (Appleton, 1954).

They are produced by the interaction between an eastward electric field ($\mathbf{E}$) in the E region, and the northward Earth's magnetic field ($\mathbf{B}$). The resulting vertical force $\mathbf{E} \times \mathbf{B}$, as shown in Figure 3-4, pushes the ionospheric plasma to move along the vertical direction. After

being lifted, the plasma is pushed downward along the magnetic field by the pressure of the above layers and by the gravity force.

This process, indicated as the Fountain Effect, is responsible for the mentioned equatorial anomalies. In these regions, the ionospheric activity usually starts from 9:00-10:00 Local Time (LT), showing its main peak around 14:00-15:00 and a secondary peak around 21:00 LT. Moreover, the equatorial anomaly shows a strong dependence on the season with severe seasonal peaks in the winter season.

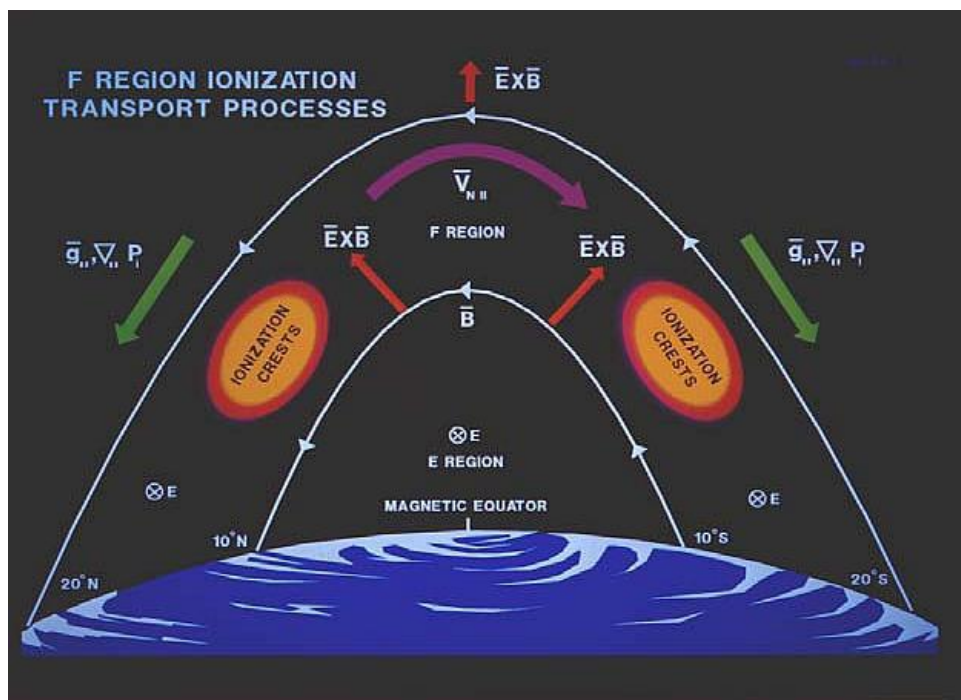


Figure 3-4: Schematic representation of the Fountain Effect responsible of the equatorial anomalies (extracted from Keith, 2009).

3.1.1.2 High latitude region

This region stretches from about 55° to 90° above and below the geomagnetic equator, and, according to the structure of the geomagnetic field (Hunsucker and Hargreaves, 2003), it can be further divided in two sub regions, namely the auroral ovals and the polar caps stretching, respectively, from about 55° to 70° and above 70° at the northern and southern latitudes. The auroral ovals are characterised by closed magnetic lines, while at the polar caps the magnetic lines are open to the solar wind (Hunsucker and Hargreaves, 2003). While at other latitudes, the main ionisation source is the Sun's radiation, at high latitudes the dominant dynamic force is represented by the interaction between high-energy particles of the solar wind with neutral gases (Hunsucker and Hargreaves, 2003).

As represented in Figure 3-5, the solar wind interacts with the Earth's magnetic field producing a bow shock which deflects the solar wind into the so termed magnetosheath. This layer interacting with the geomagnetic field generates the magnetopause which acts as a shell between the geomagnetic field and the solar wind's magnetic field, also termed as Interplanetary Magnetic Field (IMF).

Despite the shelter action of the magnetopause, part of the IMF penetrates and interacts with the geomagnetic field. At polar caps the IMF connects to the geomagnetic field producing the mentioned open field lines in contrast with the closed field lines of the auroral ovals. The polar caps' open magnetic fields allow the solar wind to move directly into the polar caps, producing large patches of increased electron density (El Gizawi, 2003).

At the auroral ovals, solar energy particles are trapped by the magnetic field and pushed by the Lorentz force to move along the closed magnetic field lines, which are vertical at these latitudes. When these particles interact with the ionosphere, they lead to the formation of small-scale irregularities. The latter are responsible also for optical and Ultra Violet (UV) emission, defined as aurora borealis and occurring between 65° and 75° degrees above the geomagnetic equator. The auroral rings extension and characteristics are strongly determined by geomagnetic disturbances.

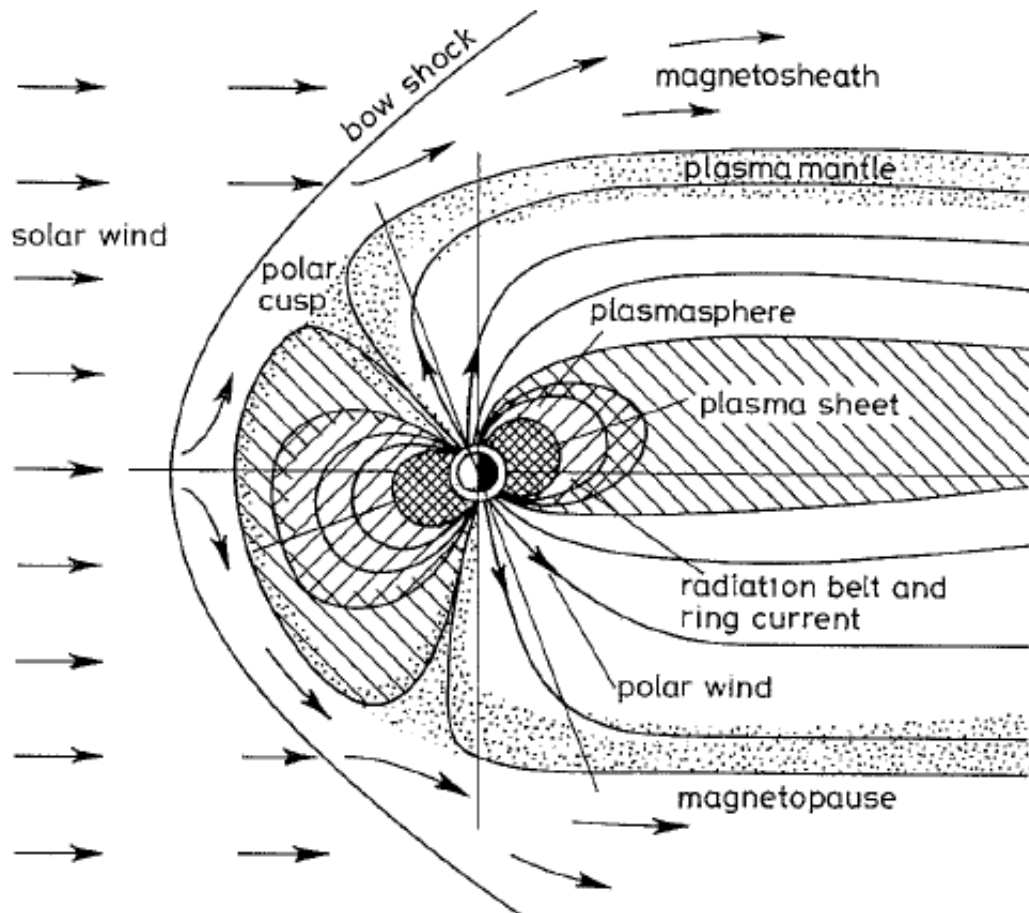


Figure 3-5: Illustration of the magnetosphere produced by the interaction between the solar wind and the geomagnetic field (extracted from Davies, 1990).

3.1.1.3 The mid-latitude region

At mid.-atitudes, extending from about 20° to 55° above and below the geomagnetic equator, the variability of the ionosphere is low and the disturbances are much less if compared with the two regions mentioned above. However, in the presence of substorms, electron densities from the equatorial regions can extend up to these latitudes.

3.1.2 Daily variations

The ionospheric daily variations are due to the 24 hours revolution of the Earth around its axis. As previously shown in Figure 3-1, because of the higher solar incidence occurring during the daytime, the electron concentration increases during the day when the photo ionisation process is dominant. On the contrary, during the night the ionisation rate is equal to zero, while the recombination process still persists, leading to a decrease in the electron density.

3.1.3 Season variatios

The ionospheric seasonal variations are produced by the revolution of the Earth around the Sun. For the D, E, and F1 layers the ionisation density increases during the summer, when the Sun's radiation is more intense, while it decreases in the winter when the radiation becomes weaker. Conversely, due to the so-termed F2 seasonal anomaly, the F2 sublayer ionisation increases during the winter (Davies,1990).

3.1.4 Solar cycle variations

As mentioned, the ionisation rate is mainly produced by sun radiations and, consequently, it changes according to the solar activity. The Sun's activity is quantified by the occurrence of sunspots, namely dark irregular areas produced by the Sun's intense magnetic activity.

The Sun follows an 11-year solar cycle. Indeed every 11 years, the Sun shows a peak on its activity, leading to the appearance of many sunspots. When there is a solar maximum, the Sun's activity strongly affects the ionosphere, producing a major number of free electrons and increasing the disturbances inside the ionosphere. As can be observed in Figure 3-6, where the sunspot number for the last six cycles is reported, the last solar peak occurred from around 2011 to 2015 at the time of this research work offering the chance to analyse the performance of the algorithms proposed in this dissertation under real severe scintillation scenarios.

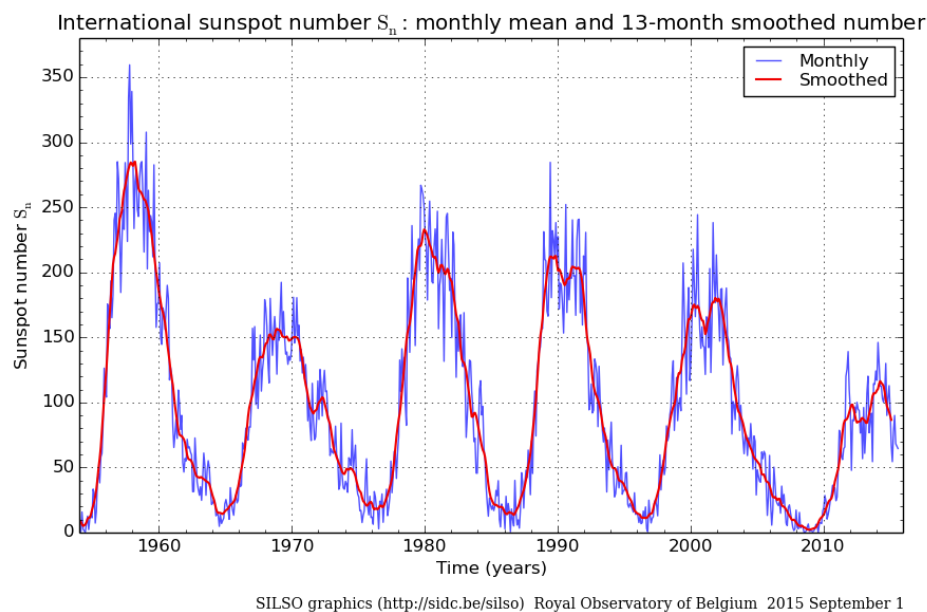


Figure 3-6: Monthly and smoother sunspot numbers for the last six cycles (extracted from <http://sidc.be/silso>).

3.2 Ionospheric scintillation

The term scintillation refers to the random signal modulation produced by structures inside the signal's propagation medium. A common and fascinating example of scintillation is given by the twinkling of the stars in the sky. The variation in the stars' brightness visible

during the night is indeed produced by the light transition through disturbances inside the atmosphere (Rino, 1979).

In a similar way, when a RF signal travels inside ionospheric irregularities, it can be subjected to random fluctuations of its phase and amplitude produced by small-scale irregularities. These irregularities can span from a few metres to a few kilometres, leading to quick variations of the ionospheric refractive index, which affects the propagation of RF signals. Although scintillation is produced mainly by diffraction, also refraction can enhance scintillation through a focusing and defocusing process (Knight, 2000; Kirtner 2009). Ionospheric signal diffraction and refraction are due to the signal group delay and phase advance produced by the presence of free electrons inside the ionosphere (Kirtner, 2009). Refraction results in a longer signal path due to the signal bending while diffraction produces the scattering of the transionospheric signal.

Specifically, signal traversing ionospheric irregularities are refracted by an amount that depends on the ionospheric Total Electron Content (TEC) and on the signal frequency. As a consequence of the refraction process the signal group velocity decreases and the phase velocity increases leading to a signal phase shift (Kintner, 2009; Kinrade, 2013) which can be quick and random in presence of ionospheric irregularities. This phenomenon strongly depends on the satellite velocity and on the ionospheric irregularity drift.

Moreover, the variations in the ionospheric refractive index are responsible for the signal scattering. Indeed, as illustrated in Figure 3-7, the signal, which has a spatially uniform phase before traversing the ionospheric irregularities, emerges from the above irregularities with a spatially irregular phase, but with a nearly constant amplitude (Kirtner,

2009). Therefore, the signal arrives at the receiver via multiple paths, and the signal on each path will interfere constructively and destructively, leading to both phase and amplitude fluctuations. The diffraction depends on the irregularity size and in fact the Fresnel zone, as explained in Appendix 1, is the fundamental length scale at which the irregularity induces scintillation. In addition, the level of scintillation is determined by the density and velocity of the irregularities.

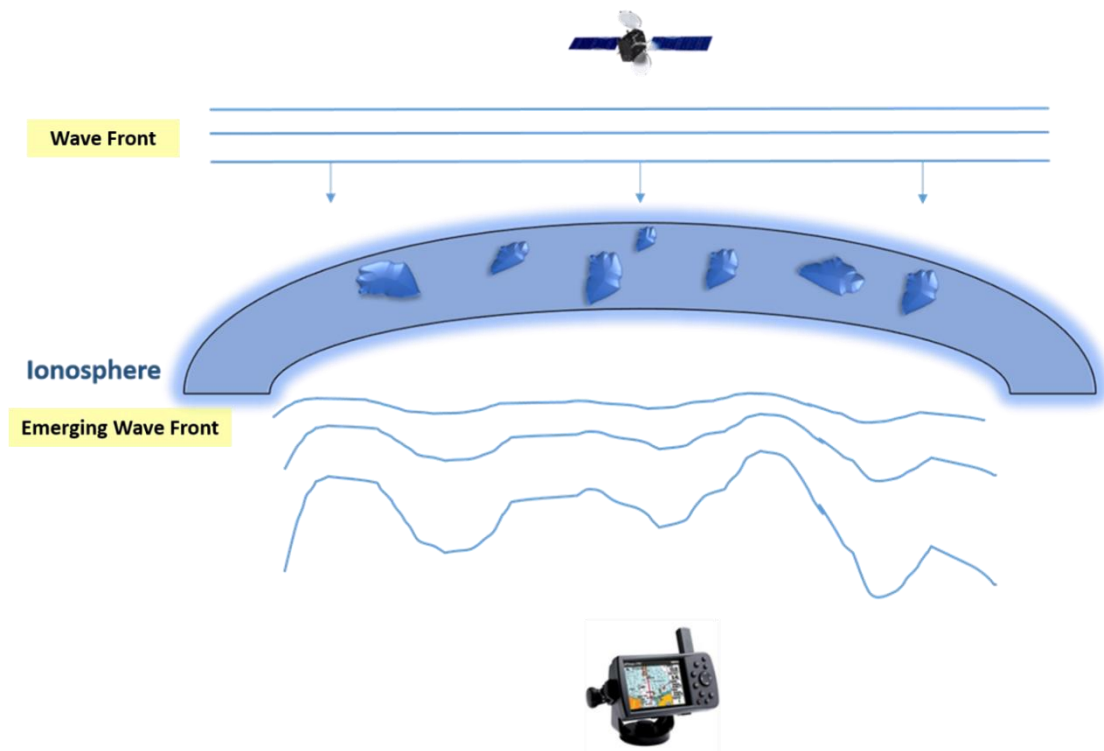


Figure 3-7: Schematic representation of the ionospheric scintillation process.

As the ionosphere's morphology is different at different latitudes, the physical processes inducing scintillation also vary, according to the geographical region of occurrence producing different effects. The worldwide distribution of ionospheric amplitude scintillation

is shown in Figure 3-8, where the regions of scintillation maximum occurrence are visible during the period of a solar maximum (left side) and a solar minimum (right side). It can be observed that the most intense scintillation activity occurs during the solar maximum at equatorial and auroral regions and at polar caps. The scintillation characteristic in the areas of maximum scintillation occurrence, namely the equatorial and high latitude regions, will be detailed in the following sections.

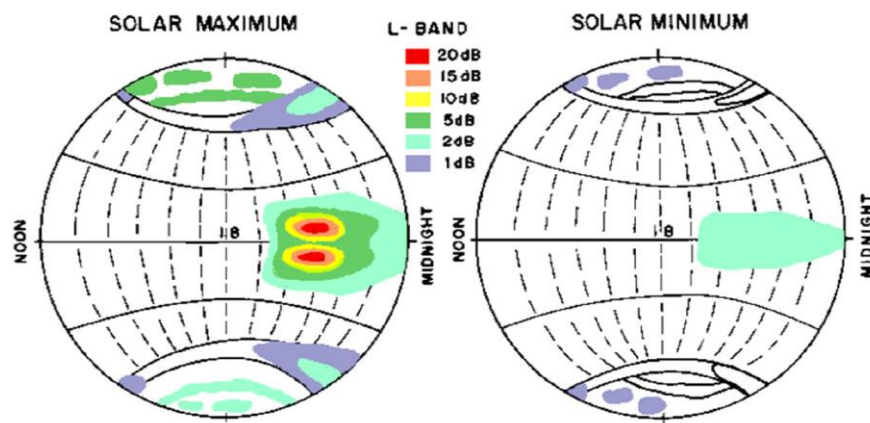


Figure 3-8: Worldwide variation of amplitude scintillation at L band (extracted from Basu et al., 1988) during a solar maximum (left side) and a solar minimum (right side).

3.2.1 Equatorial scintillation

The phenomenon is due to irregularities inside the F2 layer of the ionosphere, which arise from ionospheric bottom side instabilities (Aarons, 1982).

As explained in Section 3.1 an electric field occurs in the E region of the ionosphere. After sunset, in the F region, a further dynamo is produced. The eastward electric field described is strengthened by this new dynamo (Ghaafori, 2011) and the level of the equatorial peak is increased leading to the formation of ionospheric irregularities. The convection of the F

region produces an electron density gradient between the upper and the lower part of this ionospheric layer and induces the formation of bubbles, namely plasma depletion areas. These bubbles reach the denser upper regions and increase their dimensions, producing small irregularities at their edges, which lead to scintillation. Equatorial scintillation manifests itself with both amplitude and phase fluctuations, mainly occurring during local post sunset hours when the ionisation process rate is equal to zero and processes of instability arise inside these bubbles.

At sunrise, when the ionisation rate increases, the depletion areas are compensated and the scintillation phenomenon decreases, too.

Since equatorial scintillation is driven by the equatorial anomaly, this phenomenon increases during solar maximum periods, when the electron density is enhanced. The size of equatorial irregularities spans from a few metres to tens of kilometres of diameter (Basu et al., 1988) and the scintillation patch drift ranges from 50 to 150 km/second.

3.2.2 High latitude scintillation

At high latitude the magnetic field runs along the vertical direction, so it is in the opposite direction of the equatorial case. Consequently, the magnetic field lines (Hunsucker and Hargreaves, 2003) connect the high latitudes to the external part of the magnetosphere. While at equatorial and low latitudes the UV and X rays are the main ionisation sources at high latitudes, the interaction between the magnetosphere and ionosphere produces the main dynamic action (Tascione, 1988). Specifically, high latitude scintillation is mainly produced by geomagnetic storms, which are disturbances of the Earth's magnetosphere due to the interaction between the Earth's magnetic field and high-energy particles emitted

by the Sun. When the pressure of these particles is strong, it is able to compress the magnetosphere and interact with the magnetic field of the Earth (Ghaafori, 2011).

High latitude scintillation shows very slight daily variations while they follow seasonal and solar cycle variations, being the geomagnetic storms occurrence correlated to the solar activity. The seasonal variations are opposite to the ones observed in the case of equatorial scintillation. Indeed, high latitude scintillation is characterised by a peak in the winter time and a minimum during the summer time. Moreover, high latitude scintillation can start at any hour of the day and last from a few hours to a few days. At high latitude, the ionisation level is lower than the equatorial case. Therefore, the amplitude scintillation level is much lower than the ones occurring at equatorial latitudes. However, since at high latitudes the irregularity speed is much higher than the equatorial case (up to ten times higher) the level of phase variations is much higher, too.

3.3 Scintillation impact on GNSS receivers

Ionospheric scintillation can strongly affect the performance of a GNSS receiver and in particular its carrier tracking stage. The GPS code tracking is affected by signal amplitude variations but it is generally resistant under signal phase variation because the GPS code signal is composed by more than a thousand complete wavelengths of the GPS signal. Conversely, the carrier tracking is the receiver stage most sensitive to random signal phase and intensity variations due to ionospheric scintillation.

In fact, the deep fading induced by scintillation can make the signal intensity lower than the minimum threshold allowed for the carrier signal tracking to continue. Moreover,

random phase variations induced by phase scintillation can produce a Doppler shift, which, if wider than the PLL bandwidth, will cause the carrier tracking loop to lose signal track.

As a result an increased carrier tracking phase error, as well as navigation data bits errors, cycle slips, namely discontinuities in the carrier phase measurements, and also losses of lock can occur. When a loss of lock occurs, the signal needs to be reacquired, and, due to scintillation, the time of reacquisition could be longer than the one required in standard conditions (Seo, 2010).

As an example, in Figures 3-9 and 3-10 there are shown, respectively, the carrier to noise and Doppler frequency for GPSL1 data collected in Hanoi (Vietnam) in March 2013 and affected by both severe phase and amplitude scintillation.

It is worth underlining that when the tracking accuracy of several satellite links is decreased, the position accuracy can strongly decrease, potentially leading to navigation interruption, therefore compromising the availability and continuity of GNSS services.

One example of the above is the case of Precise Point Positioning (PPP), where the occurrence of severe scintillation can lead to errors which could be up to several metres in the vertical direction and up to tens of centimetres in the other directions. Furthermore, as explained in Chapter 4 there is an inverse relationship between amplitude and phase scintillation levels and frequency. Consequently, signals with lower frequency are more sensitive to both amplitude and phase scintillation.

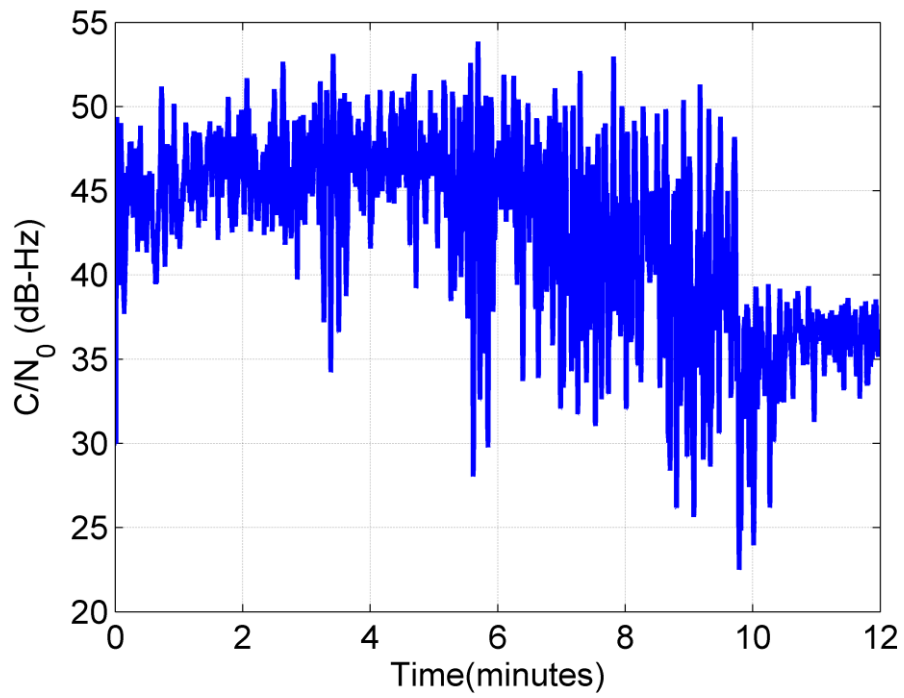


Figure 3-9: Carrier to Noise ratio for a satellite affected by strong scintillation.

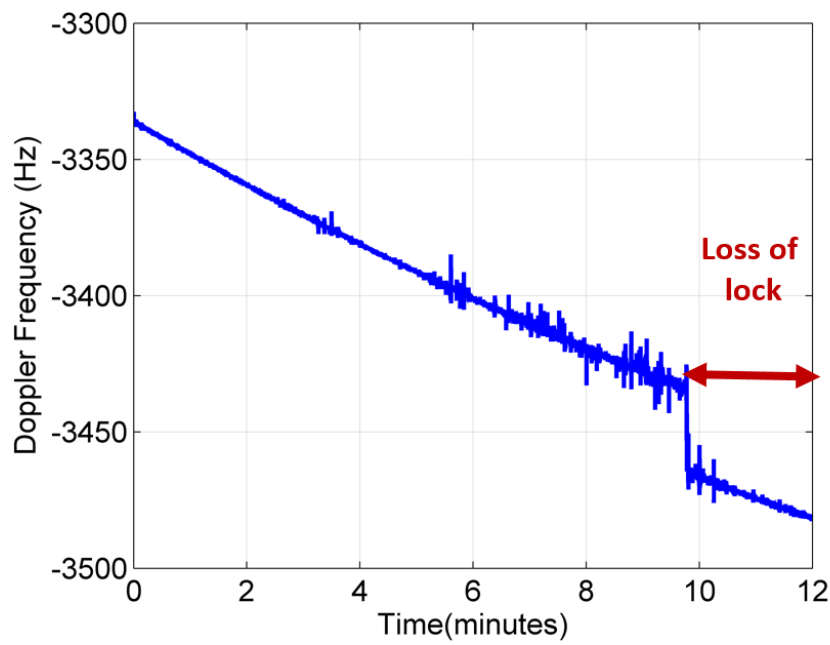


Figure 3-10: Doppler shift for a signal affected by sever phase scintillation.

3.4 Summary

This chapter provided an overview about the ionosphere morphology and on the main characteristics of ionospheric scintillation. In particular, the different physical processes driving these phenomena at different latitudes have been detailed, emphasizing that scintillation has different characteristics at different latitudes. Finally, a general description of the main effects of scintillation on GNSS receivers, and in particular on GNSS receiver tracking, has been reported. In order to quantify the level of scintillation, several key parameters can be estimated, as detailed in Chapter 4.

Chapter Four: Ionospheric Scintillation Monitoring

An ionosphere scintillation monitoring receiver (ISMR) is by definition a specialized GNSS receiver which provides information about the ionosphere and its disturbance through the estimation of a number of key parameters. This chapter has the purpose of introducing these ionospheric parameters by detailing the different algorithms necessary for their computation inside a GNSS receiver. As it will be shown, the estimation of phase scintillation parameters is strongly affected by the quality of the GNSS receiver's internal clock employed by the GNSS receiver's front end. To overcome this dependence, a technique to allow a reliable computation of the ionospheric parameters, even in the case of low quality GNSS receiver's clocks, is detailed. Finally, a number of scintillation models which can be used to assess the tracking loop of a GNSS receiver under scintillation will be introduced.

4.1 GNSS signal affected by scintillation

Section 2.4 of this dissertation introduced the model of a GNSS signal at the receiver input. When the signal is affected by ionospheric scintillation the signal model can be rewritten with the additional variation introduced in the phase and amplitude of the signal in an explicit way, as follows (Kaplan and Hegarty, 2006):

$$s_{RF}^i(t) = \sqrt{2P_i'(t)} d_i(t - \tau_i) c_i(t - \tau_i) \cos\left(2\pi(f_{RF,i} + f_D^i)t + \varphi_i'(t)\right) + n(t) \quad (4.1)$$

The differences with the expression of a generic GNSS signal expressed by equation 2.1 are given by the two following terms:

- $P'_i = P_i \delta P_i$ with P_i and δP_i representing, respectively, the nominal signal amplitude and the amplitude signal variation due to scintillation
- $\varphi'_i = \varphi_d + \varphi_s + \varphi_o$ is the phase of the received GPS carrier phase, including the contribution due to satellite and platform dynamics (φ_d), phase scintillation (φ_s) and other effects such as the oscillator noise (φ_o)

In order to evaluate the level of scintillation different parameters can be exploited, as detailed in the following.

4.2 Amplitude scintillation index

The level of amplitude scintillation affecting a GNSS signal is traditionally quantified by the amplitude scintillation index S4, which is the standard deviation of the normalized received signal intensity (SI) (Van Dierendonck et al., 1993).

This parameter is computed by exploiting the carrier tracking prompt correlator output I and Q. The latter are obtained at each time of integration T_c exploited by the I&D process block of the carrier tracking described in Chapter 2. From these raw samples the SI values can be estimated by computing the difference between the Narrow Band Power (NBP) and the Wide Band Power (WBP) over M samples (Kaplan and Hegarty, 2006) as follows:

$$WBP = \sum_{i=1}^M I_i^2 + Q_i^2 \quad (4.2)$$

$$NBP = \left(\sum_{i=1}^M I_i \right)^2 + \left(\sum_{i=1}^M Q_i \right)^2 \quad (4.3)$$

In the above equations M represents the number of integration intervals over which the above parameters are computed. For GPS L1 C/A signals, the product of M by the time of integration (T_c), expressed in ms, should be equal to 20 so that the integration is not performed over a bit transition. Then the SI can be approximated by the following:

$$SI \approx NBP - WBP \quad (4.4)$$

Therefore $S4_T$, which includes all of the environment noise components, can be estimated as:

$$S4_T = \sqrt{\frac{\langle SI^2 \rangle - \langle SI \rangle^2}{\langle SI \rangle^2}} \quad (4.5)$$

Where $\langle \cdot \rangle$ indicates the average operation performed over 60 seconds. In order to remove low frequency components due to the clock noise, multipath, troposphere and satellite geometry, and maintain the high frequency components due to scintillation, the SI sample

computed by (4.4) has to be detrended. For this purpose SI is divided by its low pass filtered version as:

$$SI_k = \frac{(NBP - WBP)_k}{(NBP - WBP)_{lpf,k}} \quad (4.6)$$

The low pass filtering is traditionally performed by using a 6th order Butterworth filter with a cut off frequency of 0.1 Hz. The rationale behind the selection of this cut-off frequency value will be discussed in Section 4.4. It is worth underlining that the low pass filtering operation at the denominator of (4.6) can introduce a delay. This implies that SI will be normalised by exploiting a delayed version of the low pass filtered SI affecting the final computation of S4. To overcome this issue a possible alternative is to replace the denominator of (4.6) with the mean of the difference between NBP and WBP computed over 60 seconds, which provides the signal trend over the time of observation, as follows:

$$SI_k = \frac{(NBP - WBP)_k}{(NBP - WBP)_{trend,k}} \quad (4.7)$$

In Figure 4-1 we see a sample case of raw and low pass filtered SI for a GPSL1 data sample collected in March 2013 in Hanoi (Vietnam) during a period of strong scintillation.

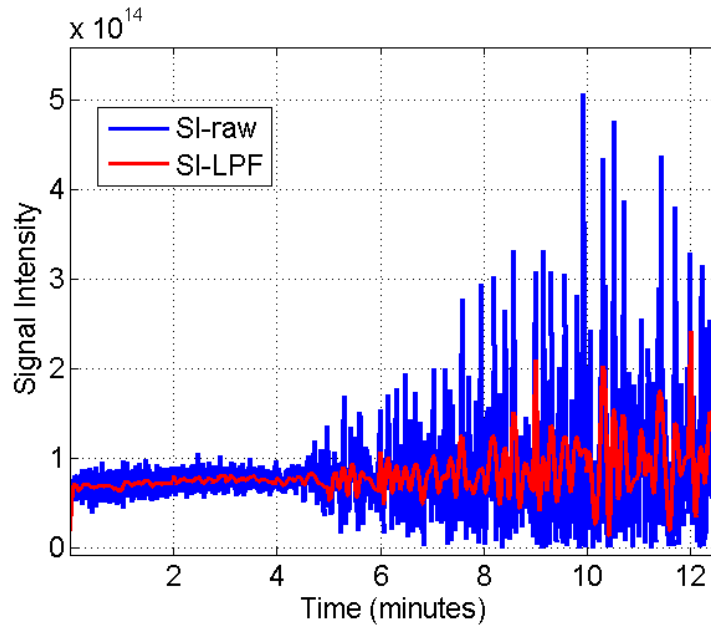


Figure 4-1: Raw and low pass filtered SI.

It is worth pointing out that the computation of $S4_T$ is affected by the environment noise given by (Van Dierendonck et al., 1993):

$$S4_{N_0} = \sqrt{\frac{100}{S/N_0} \left[1 + \frac{500}{19S/N_0} \right]} \quad (4.8)$$

Where S/N_0 is the signal carrier to noise ratio which can be computed as in (Kaplan and Hegarty, 2006). Finally, by removing the square of $S4_{N_0}$ from the square of $S4_T$, the final amplitude scintillation index $S4$ can be computed as (Van Dierendonck et al., 1993):

$$S4 = \sqrt{S4_T^2 - S4_{N_0}^2} = \sqrt{\frac{\langle SI^2 \rangle - \langle SI \rangle^2}{\langle SI \rangle^2} - \frac{100}{C/N_0} \left[1 + \frac{500}{19 C/N_0} \right]} \quad (4.9)$$

For the example shown in Figure 4-1 the total S4, including the environmental noise and the corrected S4, are reported in Figure 4-2. During the first four minutes, where the amplitude scintillation is lower the effects of the S4 correction are clearly visible.

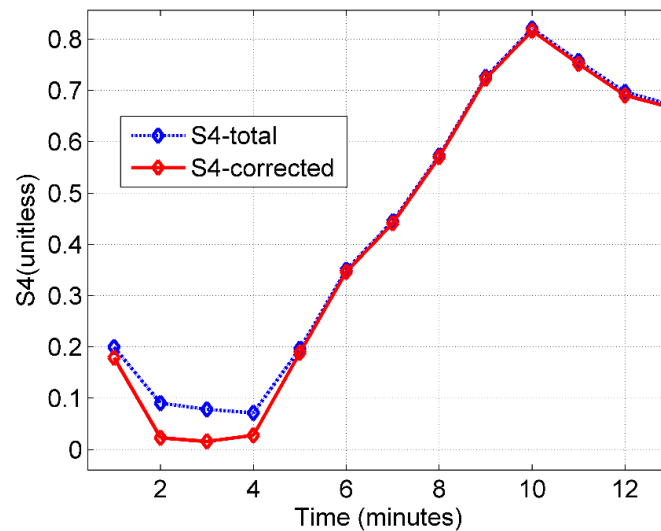


Figure 4-2: S4 before and after the correction.

In Figure 4-3 the scheme of a generic PLL including the S4 computation block is represented.

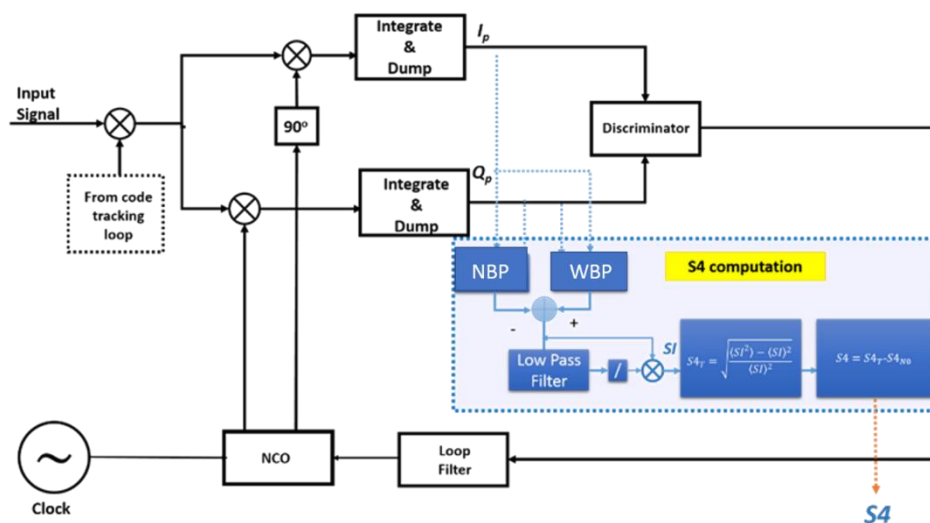


Figure 4-3: Block diagram of a generic PLL including the S4 computation block.

Finally, it should be considered that S4 is a function of the carrier frequency (Kaplan and Hegarty, 2006), as it is clear from the following relationship:

$$S4 \propto \frac{1}{f^{1.5}} \quad (4.10)$$

Consequently, for a signal with lower carrier frequencies, higher values of S4 are recorded. Indeed S4 for a GPS L2 signal is almost 1.45 times greater than the one obtained for a GPS L1 signal (Kaplan & Hegarty, 2006). Amplitude scintillation can be classified depending on the S4 value as very weak ($0.05 < S4 < 0.2$), weak ($0.2 \leq S4 < 0.4$), moderate ($0.4 \leq S4 < 0.6$) and strong ($S4 \geq 0.6$) (Dubey et al., 2006)

4.3 Phase scintillation index

To measure the level of phase scintillation the standard deviation of the detrended carrier phase σ_φ is usually exploited. In Figure 4.4 the scheme of a generic PLL including both S4 and σ_φ computation blocks is illustrated. As shown in Figure 4.4, φ is estimated by adding the phase error given by the PLL discriminator output to the PLL filter output. The sum operation has the purpose to keep the high frequencies components, which could have been suppressed by the PLL loop filter. Indeed, this happens when the phase variation range is wider than the PLL bandwidth.

Similar to the amplitude scintillation index computation, the carrier phase φ has to be detrended to remove low frequency components which are due to other sources noise different from scintillation such as the clock, multipath and also the signal Doppler shift produced by the relative motion between satellite and user.

However since the carrier phase dynamic response is much wider than the one of the signal amplitude, the detrending process is much more critical for the carrier phase case and can strongly affect the correctness of the parameter estimation (Niu, 2012).

Then the standard deviation of the detrended carrier phase can be computed by:

$$\sigma_{\varphi} = \sqrt{\langle \varphi_{HPF}^2 \rangle - \langle \varphi_{HPF} \rangle^2} \quad (4.11)$$

Where φ_{HPF} indicates the carrier phase detrended by the High Pass Filter (HPF).

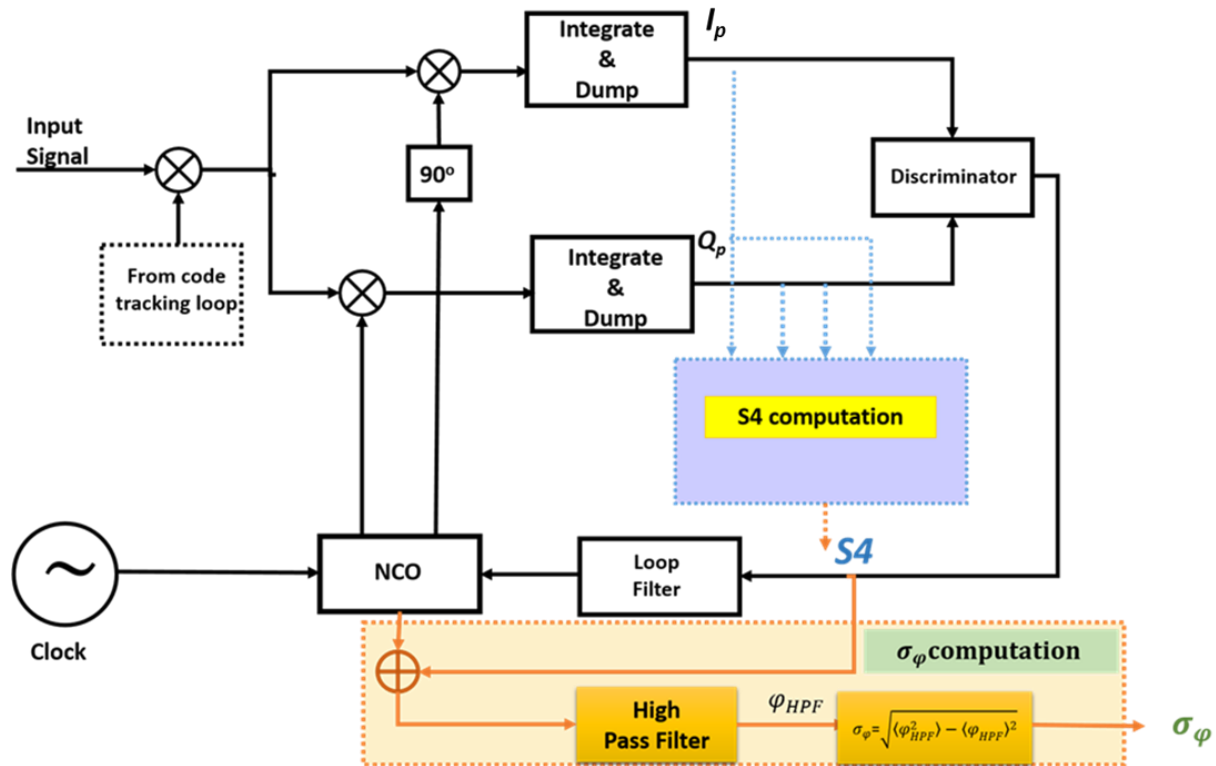


Figure 4-4: Block diagram of a generic PLL including the S4 and σ_{φ} computation blocks.

Different versions of σ_ϕ can be defined on the basis of the temporal window length used for the parameter computation. The 60 seconds version of σ_ϕ , generally indicated as Phi60, is the most used in the literature. The detrending process is conventionally performed by exploiting a 6th order high pass Butterworth filter with cut-off frequency equal to 0.1 Hz. In this case the delay introduced by the Butterworth filter does not affect the index computation as it happens with the S4 computation, since the normalisation process is not performed. Moreover the high pass filter portion, which is the one of interest for the phase scintillation index computation, is characterised by a delay almost equal to zero (Niu, 2012). However, from a practical point of view, the unique 6th order high pass Butterworth filter is usually replaced by an equivalent cascade of three 2nd order high pass Butterworth filters. This approach allows increasing the filter response stability (Niu, 2012). As sample case in Figure 4-5, the carrier phase of the data analysed in Section 4.2 is shown, while Figure 4-6 reports the same carrier phase but after the Doppler removal caused by the detrending operation. The detrended carrier phase clearly shows the strong phase fluctuations due to scintillation.

The first four minutes are missing since this is the time required by the high pass filter to converge. Finally Figure 4-7 represents the estimated Phi60 index.

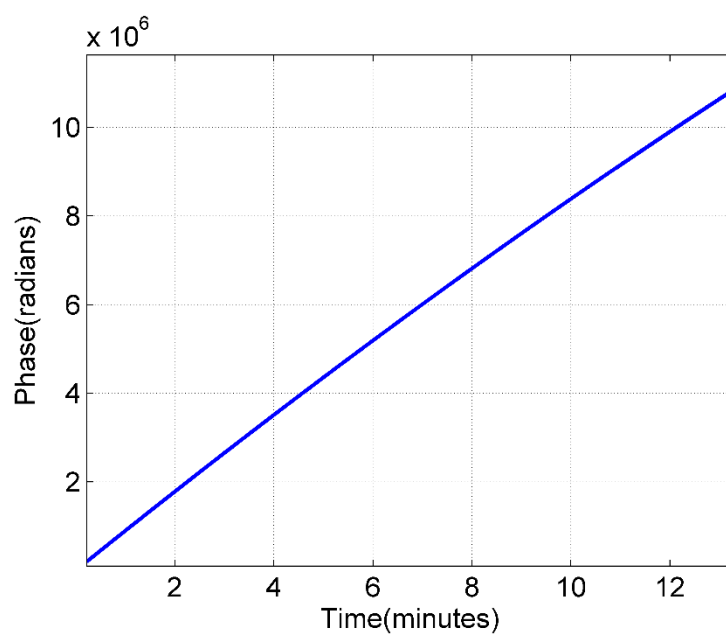


Figure 4-5: Carrier phase.

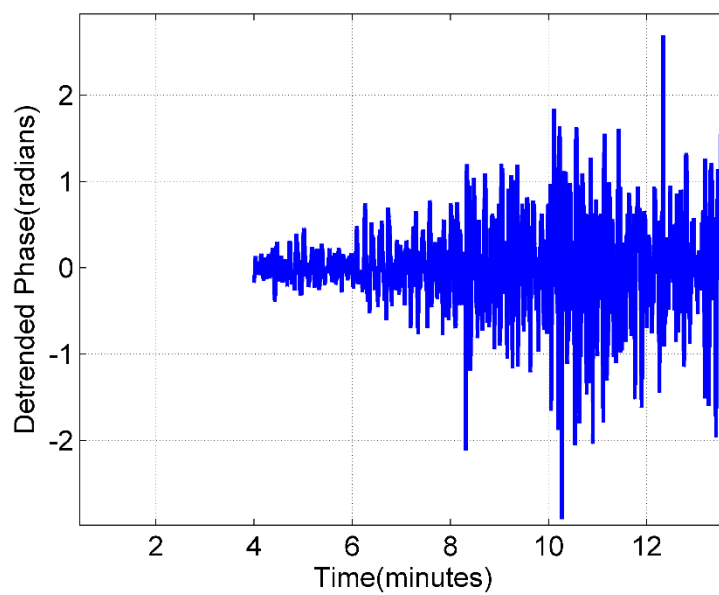


Figure 4-6: Detrended carrier phase.

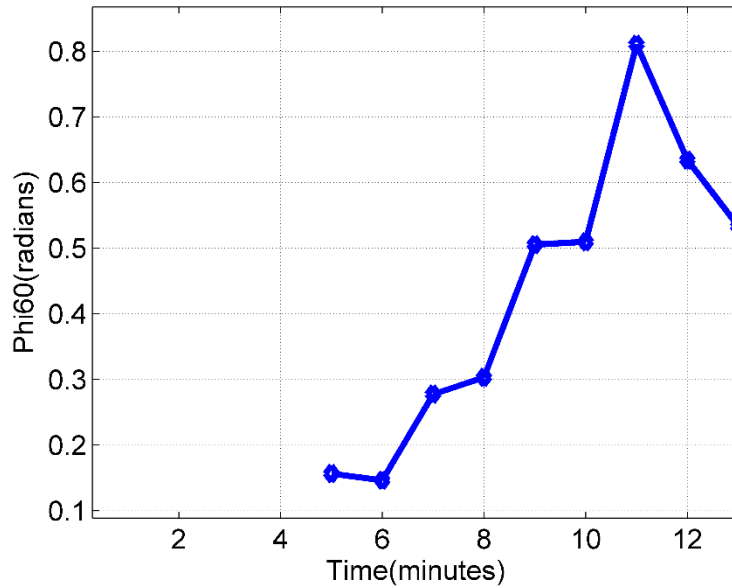


Figure 4-7: Phase scintillation index Phi60.

Although the cut-off frequency of 0.1 Hz is traditionally used by the majority of the commercial GNSS scintillation monitoring receivers, its validity has been discussed by Forte and Radicella (2002) as explained in Section 4.4. Moreover, it is worth underlining that this approach for the computation of the phase scintillation index is successful only if the clock used for the data collection is characterised by low noise. In the case of lower quality clocks, for example if using a Temperature Controlled Oscillator (TCXO) normally employed by general use GNSS receivers, a further processing of the phase error is necessary to remove the phase error introduced by the receiver's clock from the input GNSS signal phase error. Indeed, the TCXO clock noise has spectral properties similar to the ones introduced by phase scintillation (Niu, 2012) and, consequently, can lead to overrated phase scintillation evaluations. To avoid this issue, the approach proposed by Deshpande et al. (2012) can be adopted when lower quality clocks are used. It consists of

using a scintillation-free link as reference and computing the difference between the phase error of the link affected by scintillation and the one of the reference link, as in the following:

$$\tilde{\varphi} = \varphi_k - \varphi_{k_{ref}} \quad (4.12)$$

Where φ_k is the link affected by scintillation and $\varphi_{k_{ref}}$ is the scintillation-free link. The latter can be expressed as follows:

$$\varphi_k = \varphi_{k-geo} + \varphi_{clock} + \varphi_{scint} + n_k \quad (4.13)$$

$$\varphi_{k_{ref}} = \varphi_{k_{ref}-geo} + \varphi_{clock} + n_{k_{ref}} \quad (4.14)$$

In (4.13) φ_k is the sum of the phase contributions given by the satellite geometry (φ_{k-geo}), the clock (φ_{clock}), scintillation (φ_{scint}), and other noise sources (n_k). The carrier phase of the reference link ($\varphi_{k_{ref}}$) is composed of the same terms in (4.13), apart from the scintillation one, as shown by (4.14). Therefore, the differentiation process allows removing the common phase error in order to extract the scintillation one. Finally, σ_φ , similarly to S4, is also a function of the carrier frequency (Kaplan and Hegarty, 2006) and the following is valid:

$$\sigma_\varphi \propto \frac{1}{f} \quad (4.15)$$

As for S4, also σ_φ increases at lower signal frequencies. For instance, it can be estimated that the observed σ_φ for a GPS L2 signal is almost 1.28 greater than the σ_φ for a GPS L1

signal. Phase scintillation can be classified depending on the Phi60 value as very weak (0.05 <Phi60≤ 0.2), weak (0.2 <Phi60≤ 3), moderate (0.3 <Phi60≤ 0.6) and strong (Phi60 > 0.6). Finally, it is worth underlining some limitations of this index in quantifying phase fluctuations which have been pointed out by Beach (2006). Indeed Beach shows that the index is dominated by the low frequency components of the phase power spectrum which could lead to misleading values.

4.4 Scintillation spectrum

As presented before, S_4 and σ_ϕ , are variances over a fixed temporal period of, respectively, the amplitude and phase fluctuations of the GNSS signal. However, these indices do not provide information about the rate of change of signal fluctuations and, consequently, they do not give complete information on scintillation characteristics. Therefore, to fully evaluate the level of scintillation, other parameters should be evaluated. In particular, to estimate the rate of intensity and carrier phase variations the spectra of the signal intensity and carrier phase can be exploited. The Power Spectral Density (PSD) of the phase error due to scintillation can be modelled as in Rino (1979), by using the following inverse power law:

$$S_{\delta\phi}(f) = \frac{T}{(f_0^2 + f^2)^{\frac{p}{2}}} \quad (4.16)$$

Where

- T is the spectral strength of the phase noise at 1 Hz,
- p is the spectral slope of the phase PSD,
- f is the frequency of phase fluctuations and
- f_0 is the frequency of the maximum irregularity size present in the ionosphere.

Assuming that $f \gg f_0$, equation (4.16) can be approximated by $S_{\delta\varphi}(f) = Tf^{-p}$ (Conker et al., 2003), T can be computed by evaluating the strength of the detrended accumulated carrier phase at 1 Hz. The slope of the spectrum p can be estimated by computing the slope of a straight line obtained by a linear fit of the detrended carrier phase PSD, typically computed over the 0.1 to 25 Hz frequency range (Septentrio, 2016). Also, the PSD of the amplitude scintillation is inversely proportional to the frequency for high frequencies, while it can be approximated as constant at low frequencies. An example of the above is given in Figure 4-8, where the ideal phase and amplitude PSDs (in log scale) are plotted versus the frequencies (also in log scale). Moreover in Figure 4.9, as a sample case, the spectrum for a signal affected by strong scintillation is shown along with the related p and T parameters. The cut off frequency of the amplitude PSD is given by the frequency where ionospheric scintillation most contribute, indicated by the peak of the PSD. It is called Fresnel frequency f_F and can be estimated by using the following equation (Rino, 1979):

$$f_F = \frac{V_{rel}}{\sqrt{2}Z_F} \quad (4.17)$$

Where V_{rel} is the relative drift between the satellite and the ionosphere and z_F is the Fresnel zone radius which is detailed in Appendix 1. The Fresnel frequency allows identifying the frequency window where both amplitude and phase ionospheric scintillation are located (Forte and Radicella, 2002) and, consequently, is the appropriate cut off frequency to be used for the detrending process. As underlined in Sections 3-2 the ionospheric irregularity drift changes according to the latitudes and, consequently, the Fresnel frequency changes according to the geomagnetic latitude (Forte and Radicella, 2002).

Therefore, at high latitude the irregularity drift can be up to ten times higher than in the equatorial case, leading to higher values of the Fresnel frequency occur if compared to the equatorial latitude. This implies that for high latitude scenarios the cut off frequency should be higher than for the equatorial case, in order to avoid including low frequencies signal variations that are not due to phase scintillation, which could lead to the overestimation of the scintillation phase index.

Furthermore Forte and Radicella (2002) pointed out that the generally adopted 0.1 Hz cut-off frequency has been proposed for Wideband satellites which have an orbiting velocity of 3 km/sec. However, in the case of GPS satellites the Fresnel frequency should be lower due to higher orbits and, consequently, lower satellite velocities.

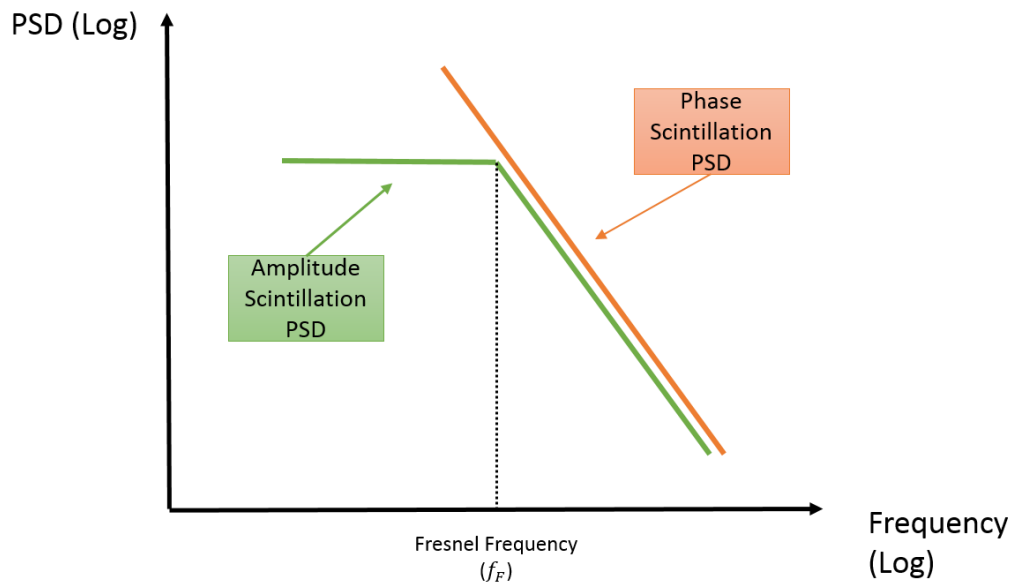


Figure 4-8: Ideal scintillation phase and amplitude PSDs versus frequencies.

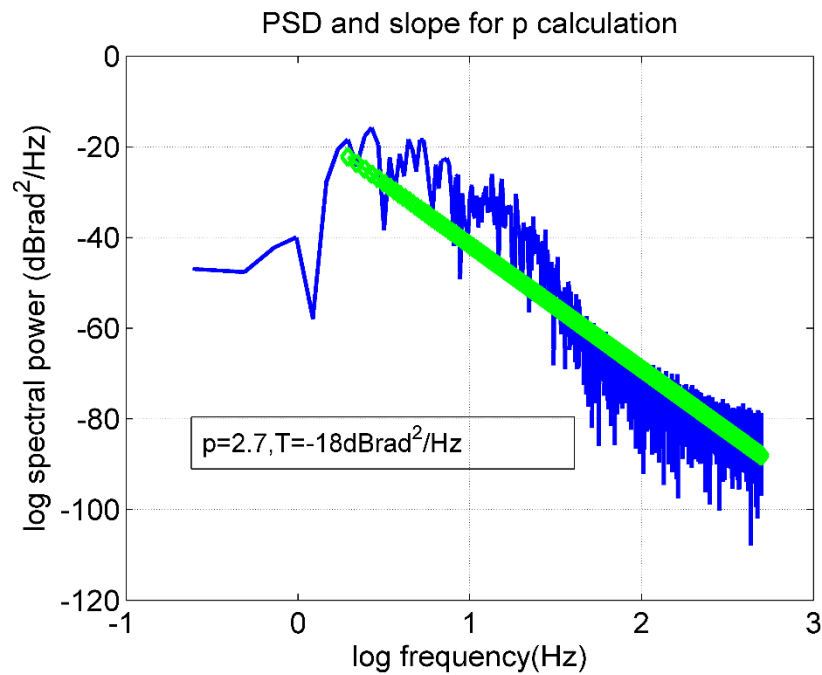


Figure 4-9: Phase scintillation Power Spectral Density (PSD) and related slope (p).

4.5 High and low latitude scintillation models

The main purpose of this dissertation is to propose improved GNSS receiver tracking algorithms resilient to ionospheric scintillation. To assess the performance of a GNSS receiver tracking algorithm under ionospheric scintillation both real and simulated data can be exploited. In order to simulate ionospheric scintillation different types of models are available. The main scintillation model categories are shortly described in the following sections.

4.5.1 Analytical models

Analytical models are constructed for specific geographical regions and therefore, they have only a local validity. They develop their signal path calculations based on the specific satellite-receiver geometry. Analytical models are constructed on empirical data and, consequently, are valid only in the regions where the stations used to collect the above data have been located. The first empirical model was proposed by Fremow and Rino back in 1973 (Fremow et al.,1973). The main limitation of this model is that it is valid only for weak scatter ionospheric scintillation conditions and, consequently, cannot be used to reproduce severe scintillation scenarios which may occur at high and equatorial latitudes. Indeed the model is based on the assumption that the autocorrelation function of the ionospheric irregularities can be approximated by a Gaussian function. Such assumption

fails in the cases of severe scintillation limiting the model application. Therefore this model is not indicated to assess GNSS receiver tracking algorithms under scintillation. Indeed for this purpose severe scintillation scenarios should be recreated to challenge as much as possible the tracking algorithms.

More advanced models, which represent a development of the above model, have been proposed to also cover high (Aarons et al., 1980), equatorial (Aarons, 1993) and low latitudes (Iyer et al., 2006). However, also these models are characterised by poor performance on a global scale.

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4.5.2 Global climatological models

Global climatological models are complementary to analytical models, being characterised by good performance on a global scale and poor performance on a regional scale. A popular global model is the Wide Band MODel (WBMOD) (Secan, 1996), which is based on the combination of a global model to reproduce ionospheric irregularities and a propagation model. The model of the ionospheric irregularities is built on libraries of data collected worldwide and includes both high and equatorial latitude scenarios. The propagation model is the phase screen model proposed by Rino (1979) The output of the WBMOD includes, among others, the amplitude and phase scintillation indices S4 and Phi60. Another popular global model is the Global Ionospheric Scintillation Model (GISM) (Beniguel, 2011). GISM exploits the Nequick model (Radicella, 2010) to reproduce the electron density inside the ionosphere and a multiple phase screens (MPS)-based scintillation model. Also, GISM provides statistical parameters of the transmitted signal

including fade durations and scintillation indices. However both models are based in the weak scattering theory and, consequently, cannot model severe scintillation scenarios.

4.5.3 In situ data-based models

In order to improve the poor performance of global climatological models on a regional scale, in situ data based (e.g. Basu et al., 1976) models exploit local data in conjunction with global models. In this way they allow achieving a wider coverage both in time and in space.

4.5.4 Statistical models

A popular statistical model is the so-called Cornell Scintillation Model (CSM) (Humphreys et al., 2008), which is mostly able to reproduce equatorial scintillation. The model is based on the assumption that the scintillation amplitude fluctuations follow a Rice distribution while the spectrum of the quickly changing components of complex scintillation is modelled by a low pass second order filter. This model can be mechanised as a simulator, which given the amplitude scintillation index S_4 and the decorrelation time as input, can provide scintillation amplitude and phase time series variations. The time series of amplitude and phase variations can be input to a hardware simulator to simulate signals affected by scintillation. Consequently, the model is particularly indicated to assess GNSS receiver tracking performance. However the main limitation is that it is not able to reproduce high latitude scintillation and very severe scintillation scenarios.

4.5.5 St. Petersburg Leeds Newcastle (SPLN) Simulator

This model is based on the combination of a complex phase method and the random screen technique (Gherm et al., 2005). Since the SPLN relies on a physical model which take as input the background and stochastic ionosphere components, it is then able to reproduce a great variety of scenarios including severe scintillation scenarios both at high and equatorial latitudes. This model can provide statistical characteristics, scintillation indices and time series of the phase and amplitude scintillations. Therefore, similar to the CSM the output scintillation signal amplitude and phase time series variations can be used in conjunction with a hardware signal simulator to reproduce the synthetic signals affected by scintillation. These capabilities of the SPLN model will be demonstrated in Chapter 6, since it was used to assess the algorithms proposed by this thesis.

4.5.6 Reuse of acquired signals

Another possible approach to reproduce a scintillated signal is to use high rate intensity and phase data collected in open sky during scintillation events. Indeed, as detailed in Chapter 6, the intensity and phase variations can be extracted from the real data through a detrending process able to isolate the high frequencies produced by scintillation. Then these intensity and phase fluctuations can be input to a hardware signal simulator to modify the intensity and phase of the generated signals. Also, this approach will be used to assess the tracking algorithms proposed by this thesis, as described in Chapter 6.

4.6 Summary

This chapter described the algorithms used by ionospheric scintillation monitoring systems based on GNSS receivers to measure the level of both amplitude and phase scintillation. Apart from the classical phase and amplitude scintillation indices S4 and Φ_{60} , it has been pointed out that the scintillation spectrum parameters should also be estimated to fully characterise ionospheric irregularities. Finally, different types of models that can reproduce scintillation scenarios have been briefly described. In particular, the SPLN model will be used to assess the performance of the algorithms proposed in this piece of work under high latitude ionospheric scintillation.

Chapter Five: Robust GNSS Receiver Tracking Loop Design

In order to cope with the variable tracking conditions imposed by scintillation events, particular attention has to be dedicated to the design of the receiver carrier tracking loop, which is the receiver stage that is most sensitive to scintillation effects. Traditional carrier phase techniques impose a trade-off between noise reduction and agility in the tracking of high dynamics. A narrow PLL bandwidth is required to reject the noise produced by scintillation fading while a wider PLL bandwidth is required to follow the fast signal dynamics induced by phase scintillation. To solve the above trade off, advanced carrier tracking architectures are proposed by the literature. For example, adaptive techniques can be exploited to select the optimum bandwidth according to the detected intensity levels and signal. This chapter first gives an overview of the receiver PLL linear model focusing on the concept of PLL equivalent bandwidth, which is a key design parameter to be considered under harsh signal tracking scenarios, as in presence of scintillation. Then the chapter reviews different carrier tracking approaches proposed by the literature to cope with scintillation events. Also, it is pointed out how the literature has shown that KF based tracking schemes represent a suitable solution to cope with scintillation events. However, these schemes could give sub-optimum solutions or also diverge in presence of severe amplitude and phase variations of the tracked signal. For this reason, it is fundamental to properly tune the KF taking into account the errors introduced by scintillation.

To this concern, the chapter proposes two novel KF based tracking algorithms, indicated as Scintillation Adaptive based KFs (SAKF1 and SAKF2), which self-tune their covariance matrices according to the detected level of scintillation.

5.1 PLL Linear Model

Due to the use of the discriminator, the intrinsic nature of a PLL is non-linear. Indeed, as introduced in Section 2.5.3, the discriminator is a highly non-linear device, which provides the phase error estimate usually by exploiting a non-linear odd function. However, in locked conditions, and consequently under the assumption of a small phase error, the tracking loop can be considered linear in its operation (Gardner, 2005) and the non-linear discriminator can be replaced by a constant discriminator gain. In this way the system can be analysed by exploiting standard linear techniques for tracking performance analysis, which allow avoiding the complex mathematical processes required to analyse non-linear systems. However, it should be considered that when the phase error increases the linearity assumption is not valid anymore and a loss of lock occurs. The graphical representation of a linear digital PLL model, obtained by replacing the discriminator with a constant gain, is reported in Figure 5-1.

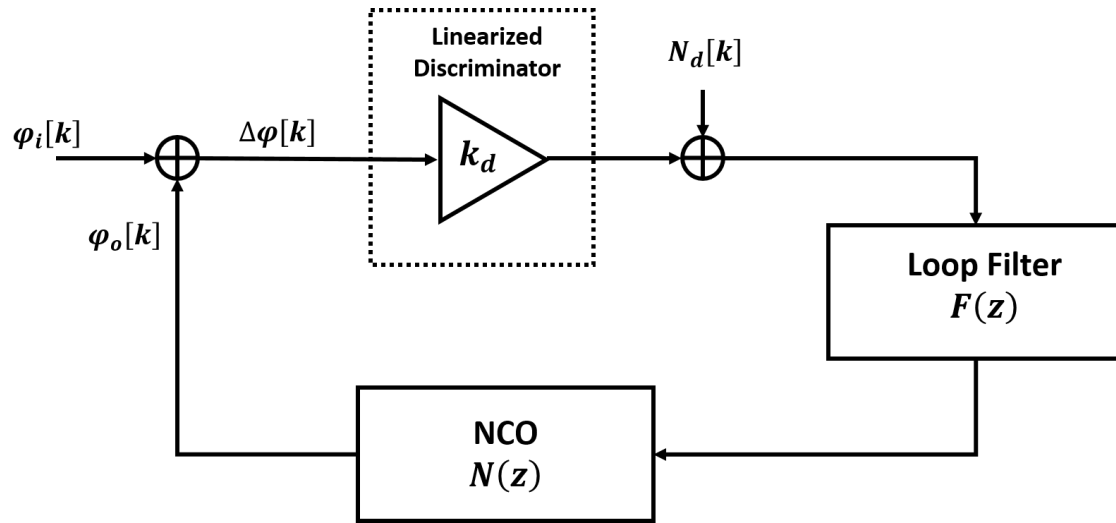


Figure 5-1: Linear PLL model.

Where

- k is the digital time
- φ_i is the input signal phase
- φ_o is the signal replica of the phase
- $\Delta\varphi$ is the phase error
- k_d is the discriminator gain obtained after the discriminator linearization
- N_d is the noise at the discriminator output which includes the signal input noise plus the noise introduced by the discriminator non linearity. This term is conventionally modelled as a White Gaussian Noise Sequence.
- $F(z)$ is the z transform of the loop filter transfer function
- $N(z)$ is the z transform of the NCO transfer function

Under the linear assumption the phase error $\Delta\varphi$ can be expressed by the following

$$f(\Delta\varphi) = k_d(\Delta\varphi) + N_d \quad (5.1)$$

Where k_d indicates the discriminator gain representing the slope of the discriminator function when the phase error is around zero (Sokolova, 2009).

The gain value changes according to the discriminator. Assuming that the discriminator gain is normalized k_d can be set equal to 1 so that the loop filter can be designed independently from the discriminator (Sokolova, 2009). Furthermore, the above linear model is updated at a rate equal to the inverse of the time of integration T_c and, consequently, φ_i , φ_o and N_d are measurements averaged over T_c .

Given the linear model assumption the new phase estimate can be expressed as:

$$\varphi_o = N(z) \cdot F(z) \cdot [k_d \cdot (\Delta\varphi(z)) + N_d(z)] \quad (5.2)$$

Consequently, the closed transfer function, namely the function that describes the relationship between the input and the output phase, of the above linear model in z domain can be so defined:

$$H(z) = \frac{\varphi_o}{\varphi_i} = \frac{k_d F(z) N(z)}{1 + k_d F(z) N(z)} \quad (5.3)$$

Different types of NCO are proposed by the literature (Kaplan and Hegarty, 2006). Assuming a phase rate feedback the transfer function of the NCO, which is a digital integrator, can be expressed as

$$N(z) = \frac{zT_c}{z-1} \quad (5.4)$$

While the loop transfer function can be defined as

$$F(z) = \sum_{k=0}^K A_k \left(\frac{T_c z}{z-1} \right)^k \quad (5.5)$$

Where k is the order of the filter and A_k are the filter coefficients. The order of the filter determines the ability in following the signal dynamics. If $k = 0$, the system transfer function $F(z)$ does not have any pole. This means that the filter does not have memory, or in other words, the filter does not take into account the past samples to estimate the new ones. Consequently, in this case $F(z)$ is a constant, which produces a first order PLL sensitive to velocity stress. Bigger values of k correspond to more poles in the transfer function, namely more memory blocks.

For instance, for $k = 1$ $F(z)$ is a linear function and the PLL is of second order which is sensitive to acceleration stress. For $k = 2$ $F(z)$ is parabolic and the PLL is of a third order which is sensitive to jerk stress. Higher orders of k are generally not used because of stability issues. Moreover, for the typical signal dynamics of a GNSS signal, values of the order minor or equal to three are enough (Curran, 2012).

When scintillation is present a third order PLL is conventionally selected in order to track the possible fast signal dynamics due to rapid signal fluctuations. Moreover, under linear assumption, the PLL can be approximated by an equivalent linear filter with noise bandwidth B_n . The equivalent noise bandwidth of a linear system with transfer function $H(j2\pi f)$ is the bandwidth of a rectangular low pass filter with the same transfer function. Therefore the loop equivalent bandwidth can be estimated by the loop transfer function as follows:

$$B_n = \frac{1}{2} \int_{0.5/T_c}^{0.5/T_c} |H(e^{j2\pi f T_c})|^2 df \quad (5.6)$$

The concept of equivalent noise bandwidth, and its relationship with the loop filter bandwidth, is illustrated in Figure 5-2.

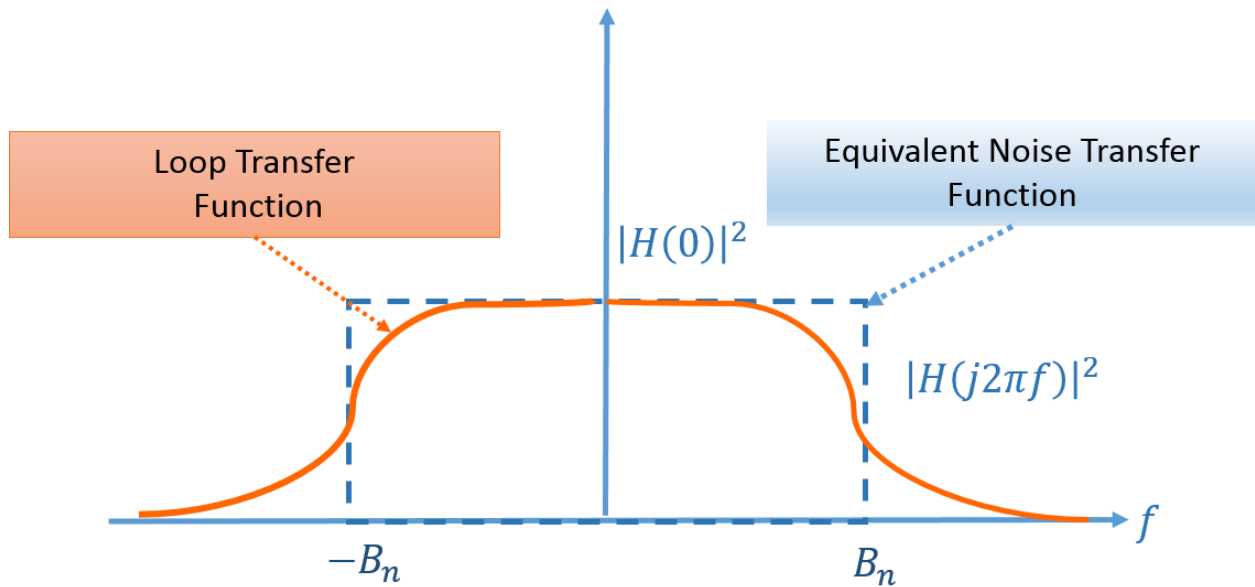


Figure 5-2: Illustration of the system noise equivalent bandwidth and loop filter bandwidth.

Furthermore, as shown in (Van Dierendonck, 1992; Ward 1998) the variance of N_d , when an *arctangent* discriminator is used, can be approximated by

$$Var\{N_d[k]\} = \sigma_n^2 = \frac{1}{2 \cdot T_c \cdot C / N_0} \left(1 + \frac{1}{2 \cdot T_c \cdot C / N_0} \right) \quad [rad^2] \quad (5.7)$$

Therefore, bigger values of T_c allow reducing the noise but at a cost of reduced dynamic tracking capabilities.

Moreover, from equation (5.6) it is clear that the noise equivalent bandwidth and order are determined by the loop filter bandwidth and order. In fact, the preferred equivalent noise bandwidth can be obtained by selecting the loop filter gains and order. Therefore, the PLL filter order and bandwidth determine the main properties of the entire PLL such as the transient response and the steady state tracking error. For this reason the loop filter is one of the fundamental blocks to be taken into account to design a PLL. More specifically, the filter noise bandwidth determines the amount of noise transferred from the input to the estimated phase output. Narrower filter loop bandwidth values enable the rejection of a larger amount of the noise and are, consequently, indicated to cope with deep fading produced by amplitude scintillation. On the other side, narrow bandwidths do not allow following fast signal dynamics produced by phase scintillation.

Scintillation scenarios can be characterized, especially at equatorial latitudes, by simultaneous variations in the signal phase and amplitude making the selection of the optimum bandwidth, or time of integration, for the GNSS signal subjected to a specific ionospheric scintillation scenario, a challenge. For this reason, as explained in the

following section, advanced tracking schemes have been proposed by the literature to cope with scintillation.

5.2 Advanced carrier tracking schemes

In Section 5.1 the basic scheme of a traditional linear PLL model is presented underlining its limitations in presence of scintillation or more in general, in the case of harsh signal tracking scenarios, such as inducing deep signal fading and/or fast signal dynamics. As already mentioned in Chapter 1, in order to overcome the above limitations a number of advanced carrier tracking schemes have been proposed by the literature. In this section, some of these architectures are briefly described.

5.2.1 FLL assisted PLL tracking schemes

As pointed out in Section 1.1, an FLL is more robust to scintillation effects than a PLL. However, frequency measurements are less accurate than the carrier ones. The FLL assisted PLL, proposed in Ward (1998), allows combining the capabilities of these two tracking schemes enjoying the advantages of both architectures.

An FLL assisted PLL tracking scheme represented in Figure 5-3 is based on the parallel combination of an FLL and a PLL. For high dynamic scenarios a third order PLL is usually combined with a second order FLL. Indeed, as pointed out in Section 5.1 a third order PLL is more robust in presence of fast signal dynamics.

However, as underlined by Chiou et al. (2007), the use of an FLL for long periods could produce losses of lock due to the noise of the frequency measurements. With this in mind, they proposed the use of an FLL assisted PLL where the FLL loop is exploited in

“backup” mode. In more detail, the FLL starts working when the signal to noise ratio is below a specific threshold and then switches on the PLL. However, under deep fading induced by scintillation, this tracking scheme switches back on the FLL tracking loop assuming that the PLL measurements are not reliable anymore. Therefore, when the nominal conditions are restored the scheme switches the PLL back on again allowing more accurate measurements.

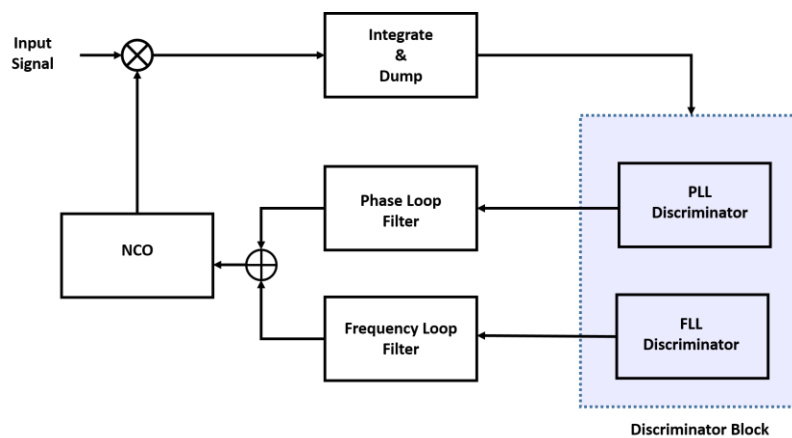


Figure 5-3: General scheme of an FLL assisted PLL.

5.2.2 Adaptive PLL schemes

In order to adapt the PLL to variable signal tracking conditions, such as those induced by scintillation, different approaches can be taken according to the parameters to be tuned and the rule exploited to adapt the above parameters.

For example, the time of integration or the loop bandwidth can be adapted according to pre-defined criteria. Alternatively, real time estimates of signal dynamics and carrier to noise density can be exploited to tune PLL bandwidth in real time, as shown by the general

scheme of an adaptive bandwidth PLL represented in Figure 5-4. Different versions of this algorithm are proposed in the literature.

For instance, the Fast Adaptive Bandwidth (FAB) (Legrand et al., 2000; Skone et al., 2005) algorithm evaluates in real time the optimal position of the loop transfer function's poles. The optimum position is the one that for a specific steady state error, due to the signal's dynamic, minimizes the thermal noise. In fact, according to the controlled root formulation (Gardner, 2005), the poles of the loop transfer function are constrained to lie on a geometric locus parameterized on the basis of a control parameter. The latter determines the decay rate of the loop impulse response. Subsequently, the FAB algorithm tunes the control parameter to meet the bandwidth requirements.

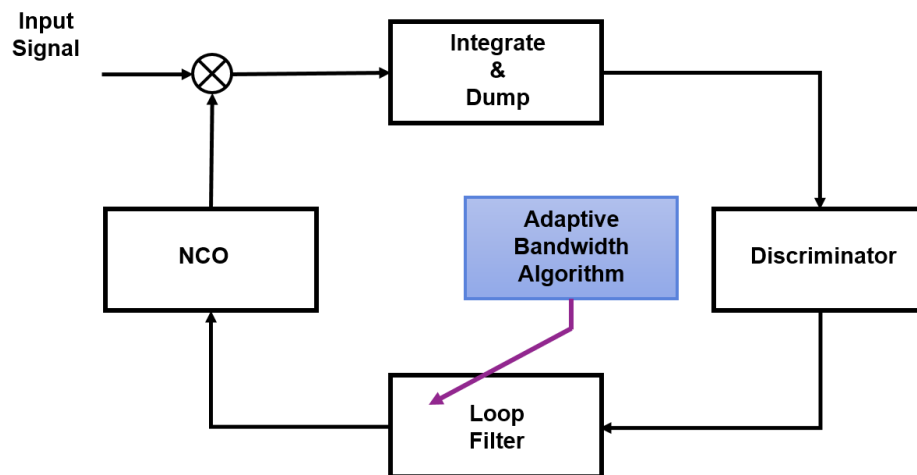


Figure 5-4: Adaptive PLL tracking scheme.

Among the adaptive PLL algorithms, the KF based PLL is an attractive solution since by definition, a KF is an optimum filter which allows minimizing the mean squared error of the parameter to be tracked (Brown and Hwang, 1997).

Different approaches can be taken in designing a KF based PLL. For instance, the KF can be used to replace only the loop filter or both the discriminator and the loop filter. Since this research work focuses on the design of KF PLL schemes optimized for scintillation scenarios, more details about this type of PLL scheme are provided in Section 5.4. KF offers several advantages under harsh tracking scenarios when carrier phase information is continuously required. Indeed, a KF allows minimizing the Mean Square Error (MSE) of the tracking filter by exploiting a dynamic and a statistical model to predict and estimate the parameters of interest representing the filter states. Moreover, it is possible to track all signal parameters by using a unique architecture instead of separating the DLL and PLL. Indeed a unique KF can replace the code and carrier loop discriminator and/or filters for each satellite channel, when a scalar tracking loop approach is used (O'Driscoll et al. 2011).

5.2.3 Vector tracking schemes

Vector tracking schemes combine the tracking and the navigation stages by considering that all the received signals are related by the configuration satellite-user. Considering all signals together it is possible to exploit the strongest signals in order to aid the weakest ones. Consequently, this tracking scheme can be very appropriate for scintillation scenarios when, usually, only some signals coming from a specific sky sector are affected by scintillation (Petovello and Lachapelle, 2006). The general structure of a vector tracking scheme is illustrated in Figure 5.5.

One of the main advantages of this scheme is the noise reduction for all channels, which helps in preventing the system from entering the non-linear region. On the other side, one

main drawback of this scheme is that any error affecting one of the analysed channels could be propagated to the other channel reducing the final performance.

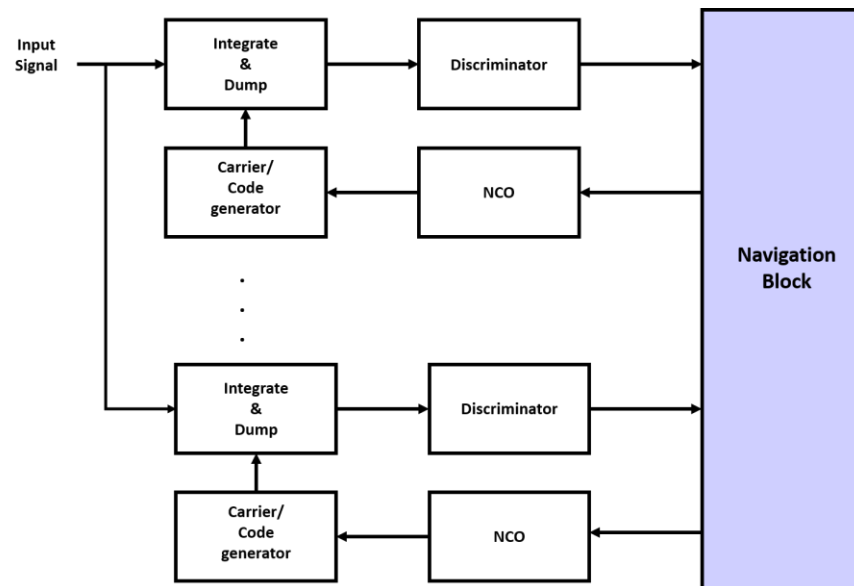


Figure 5-5: Vector tracking scheme.

5.2.4 Open loop tracking schemes

As explained in Section 2.5.3, all the closed loop architectures detailed in the previous sections maintain the signal lock by aligning the phase of the input signal with the phase of the replica signal. The alignment process is performed through a feedback system, which updates the local replica parameters in order to keep the difference between the two mentioned phases close to zero. When the signal variations are rapid and random the signal lock can be lost. In order to improve the robustness of the carrier tracking an alternative solution is represented by the open loop architecture (e.g. Curran et al., 2014).

This architecture does not rely on feedback information but on feed forward estimation of the signal parameters (Sarnadas et al., 2013) and, consequently, cannot lose the signal lock, ensuring an increased robustness. Several types of open loop architectures have been proposed in the literature (Tahir et al., 2012).

However, conventionally the majority of these schemes are based on the processing of batches of input signals at a time. Specifically, the input signal is correlated with batches of replica signals generated by using different code and frequency values extracted by a code-frequency search space. The main drawback of these algorithms is their low accuracy and high complexity. A general scheme of an open loop architecture is illustrated in Figure 5-6.

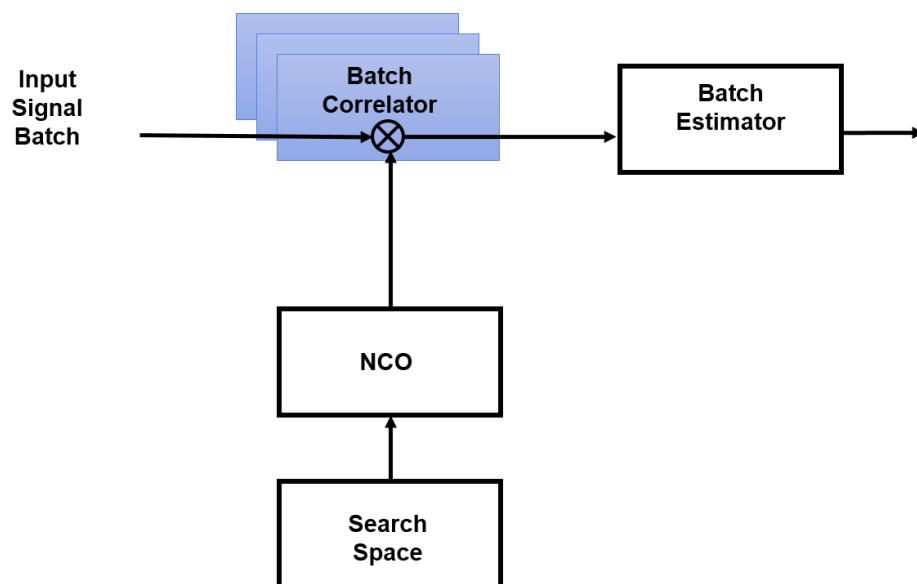


Figure 5-6: Open loop scheme.

5.3 Kalman Filter general overview

The goal of the tracking stage is to minimize the tracking error, or, in an equivalent way, the Minimum Mean Squared Error (MMSE) of the tracked parameter, namely the phase error, in the PLL case. In the case of linear approximation this can be interpreted as an optimum linear filtering problem.

In fact the objective is, when we know the spectral characteristics of an input signal, composed of the sum of a useful and a noise component, to find the optimum filter to extract the useful component by filtering out the noise. This can be accomplished by adopting different types of filters. A classical approach is to use a Wiener filter, a class of filters introduced by N. Wiener in the 1940s. The Wiener solution is essentially a weighting function, which weights the past value to determine the filter optimum output (Wiener, 1949).

The main issue with this solution is that its performance decreases in case of time variable multiple input-output problems and, also, in the case of discrete problems (Brown and Hwang, 1997). An alternative approach, which overcomes the above limitations is represented by the KF.

The latter, introduced by Kalman in the 1960s (Kalman, 1963), is a recursive solution to the problem of the linear optimum filtering which allows estimating a state of a linear system disturbed by white noise (Grewal and Andrews, 1993) through system measurements linearly dependent on the state and also corrupted by white noise. In control theory the state of a dynamic system is the minimum set of variables fully defining the system dynamic so as to enable the prediction of the system behaviour in absence of external forces (Grewal and Andrews, 1993). The filter denomination of the algorithm

proposed by Kalman is given by the property of extracting, or filtering out, the useful information from the signal affected by noise.

There is a wide literature on KF theory (i.e. Brown and Hwang, 1997; Gelb, 1974; Grewal and Andrews, 1993; Maybeck, 1994) where the details of the algorithm derivation can be found. However, for the readers' convenience, in the following the steps to implement the discrete time version of a KF are reviewed.

The basic assumption to implement a KF is that the system is linear so that the state evolution over the time can be described by the following:

$$\mathbf{x}_{k+1} = \mathbf{A}_k \mathbf{x}_k + \mathbf{w}_k \quad (5.8)$$

Where k is indicating the discrete time instant, $\mathbf{A}_k$ is the transition matrix defining the mapping function between $\mathbf{x}_{k+1}$ and $\mathbf{x}_k$, and $\mathbf{w}_k$ is the system noise which is assumed to be an additive white Gaussian Noise (AWGN) so that

$$E[\mathbf{w}_n \mathbf{w}_k^T] = \begin{cases} \mathbf{Q}_k & \text{for } n = k \\ 0 & \text{for } n \neq k \end{cases} \quad (5.9)$$

With $\mathbf{Q}_k$ indicating the covariance matrix of the system noise.

Therefore, the system dynamic model allows predicting the $(k + 1)^{th}$ state vector at the instant t_k . However, the application of a KF is also based on the assumption that measurements of the system are available and can be exploited to correct the above

prediction. The system measurement vector is considered as linearly dependent of the state vector and, consequently, can be expressed as:

$$\mathbf{z}_k = \mathbf{H}_k \mathbf{x}_k + \mathbf{v}_k \quad (5.10)$$

Where $\mathbf{z}_k$ is the measurement vector, $\mathbf{H}_k$ is the transformation matrix which maps the true state space in the observed state, $\mathbf{v}_k$ is the measurement noise vector which, as for the state measurement noise, is assumed AWGN so that

$$E[\mathbf{v}_n \mathbf{v}_k^T] = \begin{cases} \mathbf{R}_k & \text{for } n = k \\ 0 & \text{for } n \neq k \end{cases} \quad (5.11)$$

with $\mathbf{R}_k$ representing the covariance matrix of the measurement noise. Moreover, since $\mathbf{w}_k$ and $\mathbf{v}_k$ are considered uncorrelated the following relationship is valid:

$$E[\mathbf{w}_k \mathbf{v}_i^T] = 0 \text{ for all } k \text{ and } i \quad (5.12)$$

In order to implement the KF, it is also assumed that an initial estimate of the process, indicated as $\hat{\mathbf{x}}_k^-$, is available along with its covariance matrix. $\mathbf{P}_k^-$ The estimation error and its covariance matrix are indicated as:

$$\mathbf{e}_k^- = \mathbf{x}_k - \hat{\mathbf{x}}_k^- \quad (5.13)$$

$$\mathbf{P}_k^- = E[(\mathbf{x}_k - \hat{\mathbf{x}}_k^-)(\mathbf{x}_k - \hat{\mathbf{x}}_k^-)^T] \quad (5.14)$$

Where the (-) notation indicates the a priori estimates before the measurement correction process while the “hat” indicates that the parameter is an estimate.

Usually when no a priori measurements are available, due to the fact that the process mean is considered equal to zero, the initial estimates are set to zero as well (Brown and Hwang, 1997). After defining the initial parameters and defining the linear dynamic and measurement models the KF can be implemented by applying the recursive prediction and correction steps summarized by the following equations (Brown and Hwang, 1997) and graphically illustrated in Figure 5-7.

Prediction steps

- 1) Project the state ahead

$$\hat{\mathbf{x}}_{k+1}^- = \mathbf{A}_{k+1} \hat{\mathbf{x}}_k^- \quad (5.15)$$

- 2) Project the error covariance ahead

$$\mathbf{P}_{k+1}^- = \mathbf{A}_k \mathbf{P}_k \mathbf{A}_k^T + \mathbf{Q}_k \quad (5.16)$$

where $\hat{\mathbf{x}}_{k+1}^-$ is the a priori state estimate computed by projecting the state estimate through the transition matrix $\mathbf{A}_{k+1}$ and $\mathbf{P}_{k+1}^-$ is the a priori estimated covariance matrix obtained by projecting the error covariance ahead.

Correction steps

- 1) Estimate the KF gains

$$\mathbf{K}_k = \mathbf{P}_k^- \mathbf{H}_k^T [\mathbf{H}_k \mathbf{P}_k^- \mathbf{H}_k^T + \mathbf{R}_k]^{-1} \quad (5.17)$$

- 2) Update estimates with the measurements

$$\hat{\mathbf{x}}_k = \hat{\mathbf{x}}_k^- + \mathbf{K}_k [\mathbf{z}_k - \mathbf{H}_k \hat{\mathbf{x}}_k^-] \quad (5.18)$$

- 3) Update the error covariance

$$\mathbf{P}_k = [\mathbf{I} - \mathbf{K}_k \mathbf{H}_k] \mathbf{P}_k^- \quad (5.19)$$

$\mathbf{K}_k$ is the vector of the KF gains, which represent blending factors weighting the error between the real measurements and the predicted ones. The KF gains are then exploited to estimate the state a posteriori estimate $\hat{\mathbf{x}}_k$ by including the measurement $\mathbf{z}_k$ in the a priori state estimate, as shown by equation (5.17). The a posteriori estimate of the covariance matrix can be obtained from its a priori estimate by applying equation (5.19). Finally, the discrepancy between actual and estimated measurements is indicated as residual and computed as

$$\mathbf{d}_k = \mathbf{z}_k - \mathbf{H}_k \hat{\mathbf{x}}_k^- \quad (5.20)$$

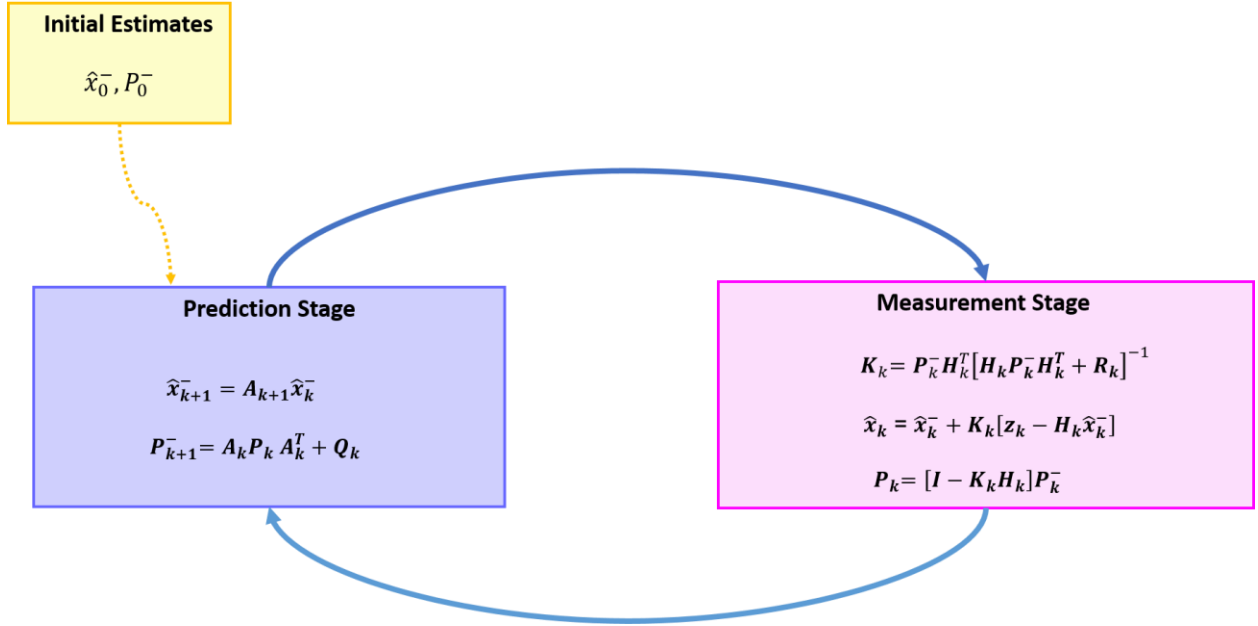


Figure 5-7: KF recursive algorithm illustration.

5.4 Linear KF based tracking loop

As introduced in Section 2.5 and repeated here for the reader's convenience, a GPS signal at the front-end output can be expressed by:

$$r_{IF}^i[k] = \sqrt{2P_i[k]} d_i[k - \tau_i] c_i[k - \tau_i] \cos(2\pi(f_{IF} + f_D^i)[k] + \varphi_i[k]) + n_i[k] \quad (5.21)$$

The signal is then processed by the acquisition stage providing a rough estimate of the initial Doppler shift and PRN code phase. After the acquisition, the tracking stage has the purpose of refining the coarse estimates of carrier phase and code-phase, also providing

an estimate of the carrier frequency. Conventional GNSS receivers generally employ two concatenated tracking loops to estimate the signal parameters. The Delay Locked Loop (DLL) is used to track delay variations while a PLL or/and an FLL allow estimating phase and frequency variations.

In this research work, to design an optimum tracking scheme, a KF has also been used to replace both DLL and PLL filters. The KF state vector (Brown and Hwang, 1997) has been defined as an error vector of four parameters. They are the code delay error $\Delta\tau$, the carrier phase error $\Delta\varphi$, the carrier Doppler frequency shift Δf_D and the Doppler frequency rate Δa . The state vector at the instant k can be represented as:

$$\mathbf{x}_k = [\Delta\tau \quad \Delta\varphi \quad \Delta f_D \quad \Delta a]_k \quad (5.22)$$

As explained in Section 5.3, once the state vector is determined, the implementation of the KF requires the definition of the system dynamic and the measurements models. Therefore, the system dynamic model is defined as in Xinhua et al. (2014)

$$\begin{bmatrix} \Delta\tau \\ \Delta\varphi \\ \Delta f_D \\ \Delta a \end{bmatrix}_{k+1} = \begin{bmatrix} 1 & 0 & \beta T_c & \frac{\beta T_c^2}{2} \\ 0 & 1 & 2\pi T_c & \pi T_c^2 \\ 0 & 0 & 1 & T_c \\ 0 & 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} \Delta\tau \\ \Delta\varphi \\ \Delta f_D \\ \Delta a \end{bmatrix}_k + \mathbf{w}_k \quad (5.23)$$

with β representing the factor used to convert cycles in units of code chips and T_c indicating the time of integration. The term $\mathbf{w}_k \sim N(0, \mathbf{Q}_k)$ is the additive zero-mean and uncorrelated Gaussian noise process vector introduced by equation (5.8).

The measurement model can be defined following two main approaches. The first methodology is to directly use the correlator output as measurements. In this case the relationship between measurements and parameters to be estimated is highly non-linear and the KF replaces both discriminators and loop filters. To cope with this non-linearity an Extended KF (EKF) should be used (O'Driscoll et al., 2011). The alternative approach, adopted here, consists in using the discriminator output as measurements so that the KF replaces only the loop filters. As pointed out above, the discriminators are non-linear devices translating the correlator outputs into state error estimates. Specifically, the carrier and code discriminators provide an estimate of the phase and code errors over the time of integration, indicated, respectively, as $\overline{\delta\varphi}$ and $\overline{\delta\tau}$ herein. The measurement model, following the approach detailed by O'Driscoll et al. (2011) is so described

$$\begin{bmatrix} \overline{\delta\varphi} \\ \overline{\delta\tau} \end{bmatrix}_{k+1} = \begin{bmatrix} 0 & 1 - \pi T_c & \frac{\pi T_c}{3} \\ 1 & 0 - \frac{\beta T_c}{2} & \frac{\beta T_c^2}{6} \end{bmatrix} \begin{bmatrix} \Delta\tau \\ \Delta\varphi \\ \Delta f_D \end{bmatrix}_{k+1} + \mathbf{v}_k \quad (5.24)$$

In equation (5.24) $\mathbf{v}_k \sim N(0, \mathbf{R}_k)$ indicates a zero-mean and uncorrelated Gaussian noise process uncorrelated with $\mathbf{w}_k$ where $\mathbf{R}_k$ is the covariance matrix of the measurement noise, as explained in Section 5.3. Assuming the code and the carrier measurements as independent, the matrix $\mathbf{R}_k$ can be represented as a diagonal matrix whose elements are the variances of the code and carrier discriminator output $\sigma_{\delta\tau}^2$ and $\sigma_{\delta\varphi}^2$.

$$R = \begin{bmatrix} \sigma_{\delta\tau}^2 & 0 \\ 0 & \sigma_{\delta\varphi}^2 \end{bmatrix} \quad (5.25)$$

For their pull-in range performance, computational cost and effectiveness at low C/N_0 (Del Pera-Rosaldo et al., 2010) an early minus late envelope and a two quadrant arctangent Costas have been selected as, respectively, code and carrier discriminators. The code and carrier measurement variances for the above discriminators are given by (O'Driscoll et al., 2011)

$$\sigma_{\delta\tau}^2 = \frac{d}{4c/n_0T_c} \left(1 + \frac{2}{(2-d)c/n_0T_c} \right) \quad (5.26)$$

$$\sigma_{\delta\varphi}^2 = \frac{1}{2c/n_0T_c} \left(1 + \frac{1}{2c/n_0T_c} \right) \quad (5.27)$$

Where d is the code delay, and c/n_0 is the carrier to noise ratio given by $10^{0.1C/N_0}$, with C/N_0 continuously computed as in Kaplan and Hegarty (2006). Once the linear dynamic and measurements models are defined the KF can be implemented by applying the recursive prediction and correction steps summarized in Section 5.3 (Brown and Hwang, 1997).

5.5 Tuning Q and R: Scintillation based Adaptive KFs

The adaptive nature of the KF is given by the variation of the gains computed by equation (5.17). If the measurements are very noisy the gains are lowered to down weight the measurements contribution, which are unreliable. Otherwise, if the measurements are not noisy the gain values increase indicating more reliable measurements. The KF equivalent filter bandwidth can be estimated by these gains as shown by Xinhua et al. (2015) and by O'Driscoll et al. (2011). Consequently, a change in the gain values corresponds to a change in the KF equivalent bandwidth. Indeed in steady state, namely when the KF gains reach constant values, the KF is equivalent to a discrete filter with same order and equivalent bandwidth.

The design of the KF requires a careful selection of the process noise covariance matrix $\mathbf{Q}$ and the measurement noise matrix $\mathbf{R}$. In a standard KF tracking architecture these matrices are fixed a priori. Traditionally, in order to define the above matrices the dominant error contributions are taken into account, as in Macchi (2010). Under the assumption of uncorrelated noise sources $\mathbf{Q}$, which is the continuous version of $\mathbf{Q}_k$ in equation (5.16), can be defined as a diagonal matrix with the diagonal elements given by the spectral densities of each process noise $[S_{\Delta\tau} \ S_{\Delta\phi} \ S_{\Delta f_D} \ S_{\Delta\alpha}]$. The relationship between continuous and discrete domain can be found in Macchi (2010). $S_{\Delta\tau}$ represents the carrier/code divergence, which can be produced by ionospheric effects. $S_{\Delta\phi}$ and $S_{\Delta f_D}$, the expected error on the phase and frequency, are conventionally considered due to the receiver oscillator error, which is assumed to be the dominant error source. The spectral densities of the clock bias and drift can be computed as (Brown and Hwang, 1997):

$$S_{\Delta\varphi} = \frac{h_0}{2} f_{L1}^2 \quad (5.28)$$

$$S_{\Delta f_D} = 2\pi^2 h_{-2} f_{L1}^2 \quad (5.29)$$

Where h_0 and h_{-2} depend on the type of oscillator and f_{L1} is the frequency of the GPSL1 signal here considered. Typical values of these parameters for different types of oscillators can be found in Brown and Hwang (1997).

$S_{\Delta a}$ is the spectral density of the expected phase acceleration which is mainly driven by the dynamics along the line of sight between receiver and satellite. For this research work the receiver is assumed to be stationary or with low dynamics, and therefore user motion and related effects were not modelled.

Even when the assumed dynamics do not correspond to the real ones due to variations in the GNSS signal working conditions, the KF may still continue the state estimation process. However, if these variations are too strong and the discrepancy between the model and the actual dynamics is too big, the KF could lead to an incorrect solution or diverge. This aspect should be considered when the KF tracking scheme is designed to be robust under scintillation. Indeed, scintillation could cause quick and random variations beyond the tolerance limit of the filter, which, consequently, may lead to the a priori fixed noise model being no longer valid.

For this reason, the algorithms proposed estimate the scintillation noise contribution and include this information in the definition of the process noise. Clearly, under severe

scintillation the ionospheric contribution should not be neglected anymore. Especially in the case of ISRMs, where low noise oscillators are generally used to capture the phase variations due to ionospheric scintillation, the ionospheric error contribution can be higher than the oscillator's. Therefore, the spectral densities of the process noise related to the phase and the frequency errors are computed as

$$S_{\Delta\varphi} = S_{\Delta\varphi_{clock}} + S_{\Delta\varphi_{scint}} \quad (5.30)$$

$$S_{\Delta f_D} = S_{\Delta f_{Dclock}} + S_{\Delta f_{Dscint}} \quad (5.31)$$

In equations (5.30) and (5.31) the sum operation is valid because the receiver's oscillator and the scintillation noise contributions are independently being generated by different physical processes. The scintillation phase noise power spectral density can be estimated as $S_{\Delta\varphi_{scint}}(f) = T f^{-p}$ by using the approximation introduced above. The frequency noise Power Spectral Density (PSD) is derived by the phase noise PSD as $S_{\Delta f_{Dscint}} = f^2 S_{\Delta\varphi_{scint}}$ (Chiou et al., 2009), where f represents the frequency corresponding to the ionospheric irregularity size, whose meaning has been explained in Section 4.4.

The proposed KF tracking schemes adapt their covariance matrix according to the working conditions determined by the level of detected scintillation. SAKF1 continuously monitors T and p and exploits these parameters to self-tune its covariance matrix.

SAKF2 first detects the level of phase scintillation by monitoring Φ_{60} , then according to the level of phase scintillation, a dynamic model is chosen among four a priori defined models. In this way it is possible to reduce the computational cost imposed by SAKF1.

The models used in SAKF2 correspond to absent, weak, moderate and severe scintillation scenarios. The $S_{\Delta\varphi}$ and $S_{\Delta f_D}$ elements in the covariance matrix for each of the above scenarios were defined by using empirical values for p and T obtained in an experimental way and reported in Table 5-1. $S_{\Delta\tau}$ and $S_{\Delta\alpha}$ are kept constant and modelled as in Macchi (2010).

Table 5-1: Phase scintillation scenario definition.

Phase Scintillation Scenarios	Φ_{60}	p	T
H0: (Scintillation free/Very weak scintillation)	$\Phi_{60} \leq 0.2$	0	0
H1: Weak Scintillation	$0.2 < \Phi_{60} \leq 0.3$	1.2	-39 dB2/Hz
H2: Moderate Scintillation	$0.3 < \Phi_{60} \leq 0.6$	3.4	-25 dB2/Hz
H3: Strong Scintillation	$\Phi_{60} > 0.6$	4	-15 dB2/Hz

In order to tune the measurement noise according to the signal intensity, the C/N_0 is estimated as in Van Dierendonck et al. (1995) and used to estimate the discriminator variance output in (5.26) and (5.27). In this way the KF gains, and consequently the loop equivalent bandwidth, change according to the signal intensity. In case of signal fading the loop noise is decreased to filter out, as much as possible, the noise affecting the signal parameter estimation. However, there could be cases when, even if C/N_0 is low, the bandwidth should not be decreased too much to follow the fast signal dynamics due to phase scintillation and, consequently, to avoid losing the signal's lock, which is the main goal of the proposed algorithms. To further enhance the robustness of the measurement noise estimation, a weighting factor is applied to the measurement noise matrix. To

estimate this factor first of all the measurement noise matrix is evaluated as suggested by Yang et al. (2012), to make the actual value of the covariance of residuals as close as possible to its theoretical value. The covariance of residuals can be defined as:

$$\bar{\mathbf{C}}_k = \frac{1}{m} \sum_i^m \mathbf{d}_k \mathbf{d}_k^T \quad (5.32)$$

Where $\mathbf{d}_k$ represents the KF residuals defined by (5.20) and m is selected equal to 10 samples.

Therefore the theoretical covariance is given by

$$\mathbf{C}_k = E\{\mathbf{d}_k \mathbf{d}_k^T\} = \mathbf{H}_k \hat{\mathbf{P}}_k^- \mathbf{H}_k^T + \mathbf{R}_k \quad (5.33)$$

Considering that the predicted innovation covariance should match the one computed by the innovation values, any discrepancy between the two covariance matrices should be attributed to a mis-modelling of $\mathbf{P}$ or/and $\mathbf{R}$. It can be shown (Yang, 2012) that, assuming $\mathbf{Q}$ as properly defined, the degree of mismatch between the theoretical and the actual covariance can be compensated by multiplying $\mathbf{R}$ by a scaling factor

$$\mathbf{R}'_k = \alpha_k \mathbf{R}_k \quad (5.34)$$

with $\alpha_k = \frac{\text{trace}(\mathbf{C}_k)}{\text{trace}(\bar{\mathbf{C}}_k)}$

The general structure of the SAKF1 and SAKF2 algorithms are reported, respectively, in Figure 5-8 and 5-9. It is shown how scintillation parameters and C/N_0 are computed by dedicated blocks, and fed into the KF replacing the traditional PLL and DLL loop filters. SAKF1 directly estimates p and T , while SAKF2 selects them from a set of a priori defined values after defining the scintillation scenario as detailed above.

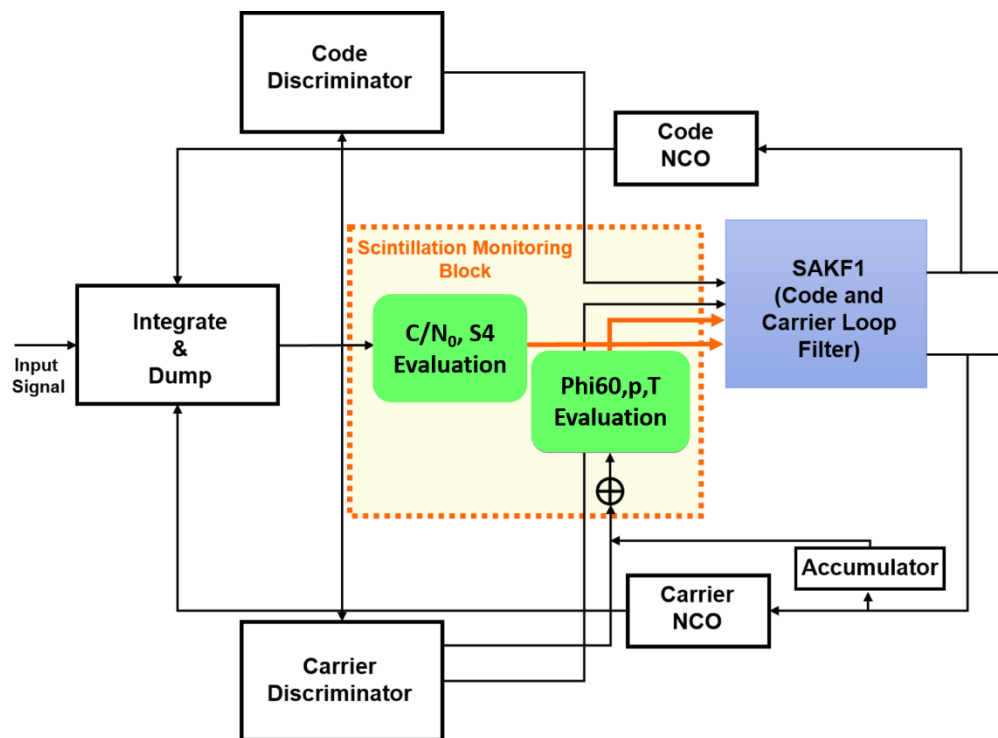


Figure 5-8: SAKF1 structure.

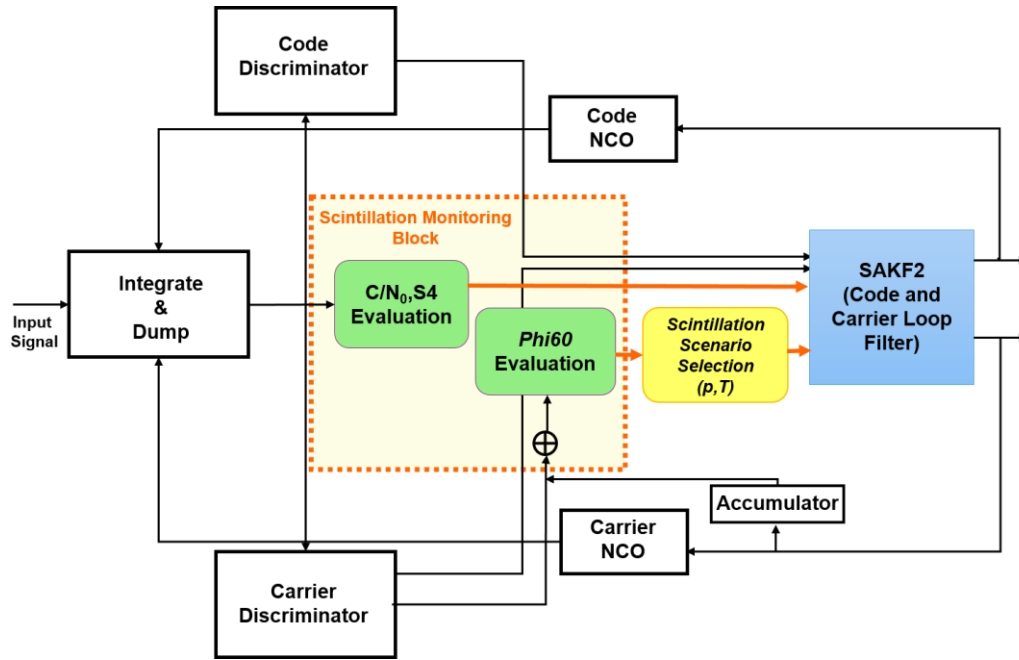


Figure 5-9: SAKF2 structure.

5.6 Summary

This chapter introduced the main limitations of a classical PLL in presence of scintillation and reviewed some of the main advanced tracking architectures proposed by the literature to cope with scintillation scenarios. Therefore, since the literature has shown that KF tracking schemes are particularly suitable to cope with scintillation effects two scintillation based adaptive KFs, SAKF1 and SAKF2, which self-tune their covariance matrix according to the detected scintillation level are proposed. SAKF2 allows reducing the computational cost of SAKF1 avoiding the continuous estimation of the scintillation spectrum parameters, p and T , and selecting the values of p and T among predefined ones, after defining the level of phase scintillation through the estimation of Φ_{60} . The

results obtained by assessing the performance of SAKF1 and SAKF2 under both simulated and real scintillation will be presented in Chapter 6.

Chapter Six: Tracking Algorithms Assessment

This chapter presents the results obtained by the assessment of the novel tracking algorithms proposed by this thesis and detailed in Chapter 5.

Specifically, the tracking performance of the **SAKF1** and **SAKF2** algorithms has been compared to that of traditional algorithms, such as PLL schemes with fixed PLL bandwidth values, classical adaptive KFs based tracking schemes and to a commercial ISMR, the Septentrio PolaRxS receiver. Since, as pointed out in Chapter 3, the scintillation characteristics, and consequently their effects on the receiver tracking stage, are different at different latitudes, the above comparison has been carried out by exploiting both high and low latitude data. Therefore, this chapter can be distinguished in two main parts.

The first part describes the experimental set up and the methodology to simulate a **high latitude scintillation scenario**. Then, it reports the results obtained by assessing the tracking algorithms under the simulated high latitude scenario. The second part of the chapter is dedicated to the assessment of the algorithms using both real and simulated **equatorial data**. Algorithms' performance are evaluated in terms of tracking and scintillation monitoring capabilities.

6.1 High Latitude Scenario: experimental set up

In this section the results obtained by assessing the proposed algorithms under a high latitude scenario are reported. Since real high latitude data were not available, the capabilities of the hardware GSS8000 Spirent signal simulator available at the University of Nottingham have been exploited in order to recreate the high latitude scintillation scenario.

The general scheme of the experimental set up used to simulate the above scenario, and to test the proposed algorithms, is illustrated in Figure 6.1. The Spirent simulator, which allows generating GNSS signals in a controlled environment, has been exploited in conjunction with the SPLN, the physics based scintillation model described in Section 4.5.5. The model has been configured in order to recreate a severe high latitude scintillation scenario with the purpose to challenge as much as possible the carrier tracking schemes under test. Afterwards the model outputs, namely the time series of scintillation amplitude and phase variations, were formatted in a user command defined (.ucd) file, a particular type of file which can be input to the Spirent signal simulator to modify its generated signals.

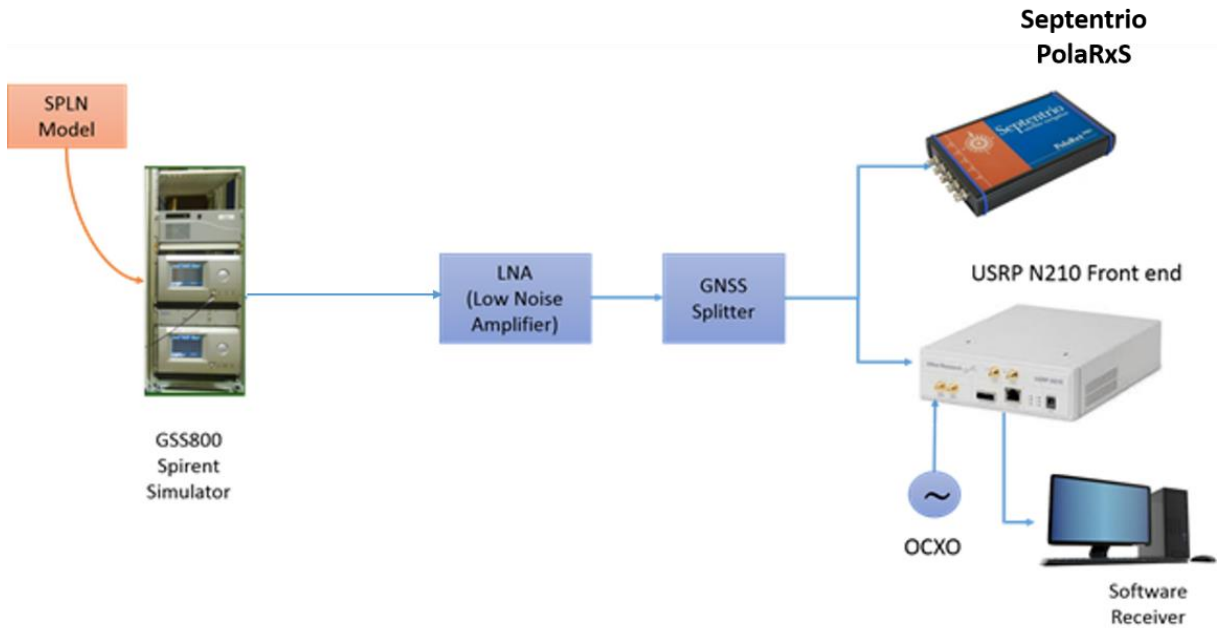


Figure 6-1: Experimental set up scheme.

It is worth pointing out that the GSS8000 Spirent signal simulator allows adding scintillation effects on the signals also by exploiting its internal statistical model, namely the Cornell Scintillation Model (Hinks et al., 2008) introduced in Section 4.5. However this model can only recreate equatorial scenarios and consequently is not suitable for the purposes of this experiment.

Once the signal affected by high latitude scintillation was produced, the RF data was collected by using a Universal Software Radio Peripheral (USRP) N210 front end by Ettus connected to the Spirent simulator. The USRP N210 is a flexible radio platform which can be configured to process any signal from DC to 5.9 GHz (Peng et al., 2012) by simply adding the daughterboard specific for the purpose. Therefore this front-end is capable to process any L band GNSS signal, even if, due to bandwidth limitations, only a sub set of

these signals can be processed at the same time. In the case of scintillation monitoring applications the choice of the front-end is crucial. Indeed, as pointed out by Peng et al. (2012), a front end used for scintillation monitoring applications should be able to minimize:

- the thermal noise
- the oscillator phase noise
- out of band interference

The USRP N210 was selected for its flexibility and since its validity for scintillation monitoring applications has been already proven by other studies (e.g. see Peng et al., 2012). However one main drawback of this front-end is the poor frequency and phase stability of its internal TCXO clock (Peng, 2012). To overcome this limitation an external low noise OCXO by Total Frequency Control (OS364-10) has been connected to the USRP.

An OCXO was selected since a low noise clock is preferable to catch, without ambiguity, the phase variation due to scintillation. Indeed the main drawback of using a TCXO is that the spectral phase fluctuations produced by the instability of this clock overlap those caused by ionospheric scintillation. Consequently, the clock phase variations could be wrongly detected as scintillation phase variations and affect the receiver scintillation monitoring capabilities. This is why, when a TCXO is used, specific algorithms, as the ones described in Section 4.3, should be implemented to identify phase fluctuations produced by scintillation. Therefore, from the above considerations the crucial importance of the clock characteristics for scintillation monitoring purposes is evident.

In order to collect the data from the front end and to configure the data collection setting, i.e. sampling frequency, signal to be captured etc., the GNU Radio Software interface has

been used. Moreover, a commercial ISMR, namely the Septentrio PolaRxS receiver, used as benchmark, has been also connected to the simulator along with the above front end through a signal splitter. The PolaRxS is a multi-constellation multi-frequency ISMR housing an ultra-low noise OCXO (Septentrio, 2014).

In Figure 6-2, a picture of the test set up is shown. Specifically, we can see the interface of the Spirent simulator and the GNU Radio Software interface used to collect the data using the USRP N210. Through the GNU Radio Software Interface, GPS L1 data has been collected, with a sampling frequency equal to 5 MHz and a bit resolution equal to 16 bit. These recording parameters were proven to be suitable for scintillation monitoring purposes by Peng (2012). Details of the data collection are summarized in Table 6.1

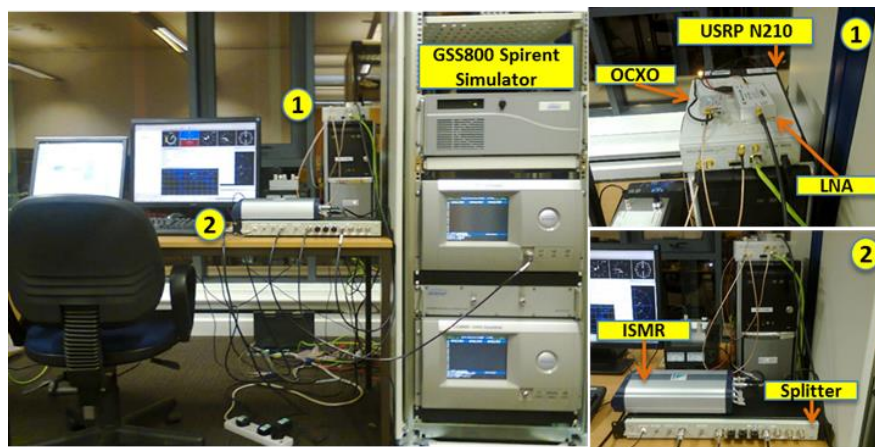


Figure 6-2: (Left) Experimental test set-up; (Right up) zoom of the USRP N210 connected to an external OCXO and to a LNA; (Right down) zoom of the splitter connecting the simulator output to a commercial ionospheric scintillation monitoring receiver (ISMR) and to the USRP N210.

Table 6-1: Data collection details.

Parameter	Value
Signal	GPS L1
IF	0
Sample Type	Complex
Bits per sample	16
Sampling Rate	5 MHz

6.1.1 Data processing methodology

The IF data collected as explained above has been post-processed by using the following different tracking architectures, implemented in a full software receiver (Borre et al., 2007).

- Conventional tracking schemes with fixed bandwidth;
- The proposed SAKF1 and SAKF2 and;
- Conventional adaptive KF tracking schemes, indicated as AKF1 and AKF2.

Appendix B provides a general scheme of the implementation of these algorithms in the above software receiver. Furthermore, the algorithms' details are reported in Table 6-2. Specifically, AKF1 is a classical adaptive KF tracking algorithm with the covariance matrix $\mathbf{Q}$ a priori fixed as in Macchi (2010) and with the scaling factor, described by equation (5.34), applied to the measurement noise matrix $\mathbf{R}$. The time of integration T_c has been set equal to 10 ms, which is the same default value adopted in the Septentrio PolaRxS receiver. Moreover, the spectral noise densities of the clock bias and drift in the KF based algorithms have been computed considering that an OCXO was used for the data collection (Brown and Hwang, 1997). AKF2 has the same characteristics as AKF1 but it exploits a time of integration equal to 20 ms. The latter value was selected to observe how much the tracking performance decreases by extending the time of integration when high dynamics are present as in the high latitude case.

Therefore, two different settings were adopted for the algorithms with fixed bandwidth. One scheme was configured so as to have a PLL bandwidth equal to 15 Hz and a time of integration equal to 10 ms. The above setting values are typically used in commercial ISMRs. The other tracking scheme was configured with a bandwidth equal to 4 Hz and a

time of integration equal to 20 ms so as to minimize the noise using a narrow bandwidth and a longer time of integration.

Table 6-2: Tracking algorithms details.

	Q	R	T _c (ms)	B (Hz)
AKF1	A priori fixed	Variable with scaling factor (eq. 5.34)	10	Not applicable
AKF2	A priori fixed	Variable with scaling factor (eq. 5.34)	20	Not applicable
SAKF1	Variable	Variable with scaling factor (eq. 5.34)	10	Not applicable
SAKF2	Variable	Variable with scaling factor (eq. 5.34)	10	Not applicable
Fixed B (T_c=20ms)	Not applicable	Not applicable	20	4
Fixed B (T_c=10ms)	Not applicable	Not applicable	10	15

6.1.2 Tracking performance evaluation criteria

Ionospheric scintillation can affect the tracking loop of a GNSS receiver by increasing the system noise, and consequently producing deep signal fading, and inducing Doppler shifts due to the signal phase fluctuations. Since, as explained in Section 5.2, there is a trade-off between noise reduction capabilities and the ability in following the signal dynamics, both these aspects are analysed to assess the algorithms' performance.

Specifically, the phase tracking performance of the above algorithms is assessed by evaluating the phase jitter, which has been computed over temporal windows of 4 seconds in order to reduce the noise affecting the measurement. The latter has been computed by exploiting In Phase (I) and Quadrature (Q) components as (Forte, 2012)

$$Phase\ Jitter = std\left(\operatorname{atan}\left(\frac{Q}{I}\right)\right) \quad (6.1)$$

Moreover the Phase Lock Indicator (PLI) has been also computed as in Van Dierendonck (1996). This tracking indicator, which is proportional to the cosine of twice the phase error, is also estimated by using the I and Q components as

$$PLI_k = \frac{1}{M} \sum_{m=0}^{M-1} \frac{I_{k-m}^2 - Q_{k-m}^2}{I_{k-m}^2 + Q_{k-m}^2} \approx \cos(2\delta\varphi) \quad (6.2)$$

Where M has been selected equal to 100 samples in order to average the noise (Bhaskar, 2015). The parameter can be used as a qualitative measure of the carrier tracking performance and it assumes values between -1 and 1. PLI values close to 1 indicate good tracking capabilities while, vice-versa, lower values indicate a degradation of the tracking performance. A value of PLI below 0.86, when, as in this case, a Costas discriminator is used, corresponds to a phase error bigger than the threshold of 15 degrees (Van Dierendonck 1996), namely the 1-sigma phase error threshold for the carrier phase tracking commonly used for commercial GNSS receivers. Finally, to compare the tracking performance in terms of agility in following the signal dynamics, the carrier Doppler and its standard deviation are compared as well. These parameters will be exploited also to assess the algorithms under equatorial scintillation scenarios, described in Section 6.2 and 6.3.

6.1.3 Test results

As explained above, the SPLN input were selected so as to recreate a very harsh high latitude scintillation scenario. In Figure 6-3, the $S4$ and Φ_{60} values recorded by the Septentrio PolaRxS ISMR are shown for the simulated data set.

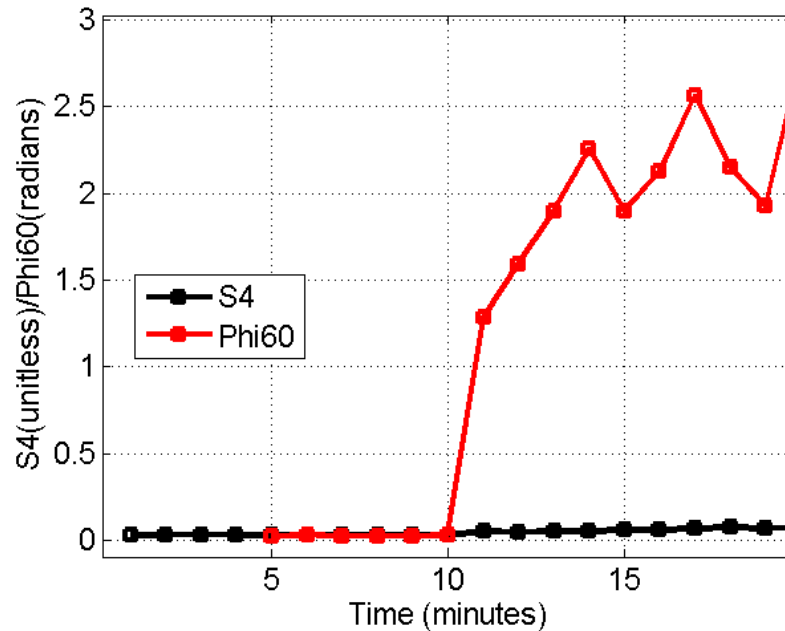


Figure 6-3: Scintillation parameters recorded by the Septentrio PolaRxS for the simulated high latitude scenario.

For Φ_{60} , the first four minutes of data are missing since this time is required by the detrending filter to converge. As expected, the values of $S4$ are at noise level since high latitude scintillation is characterized by very weak amplitude fluctuations. On the other hand, for Φ_{60} very high values are observed, since, as previously underlined, intentionally an extreme scenario has been created in order to challenge the carrier tracking algorithms.

In order to assess the carrier phase noise performance of the algorithms, the phase error jitter has been estimated for all tracking schemes, as reported in Figures 6-4 where the red line indicates the 15 degree tracking threshold. As described in Section 6.1.2, the phase jitter is computed by exploiting the I and Q correlator outputs. Since the Septentrio PolaRxS ISMR provides also the 50 Hz I and Q components, it has been possible to compute the phase jitter also for the above receiver.

It can be noticed that the Septentrio ISMR shows the best performance in terms of noise reduction in absence of scintillation. This can be attributed to the lower level of noise of the OCXO embedded into the Septentrio receiver if compared to the OCXO used to collect the IF data. However in presence of scintillation the KF based tracking schemes outperform the algorithms with fixed bandwidth, as well as the Septentrio receiver. This can be better observed in Figure 6-5 where two details of Figure 6-4, in correspondence of scintillation peaks, are shown.

Furthermore, SAKF1 and SAKF2 achieve better performance than the classical KF based tracking schemes, indicated as AKF1 and AKF2. In particular, the SAKF1 algorithm, which continuously monitor the scintillation parameters to self-tune its covariance matrix, is the only one whose phase jitter never goes above the 15 degrees threshold.

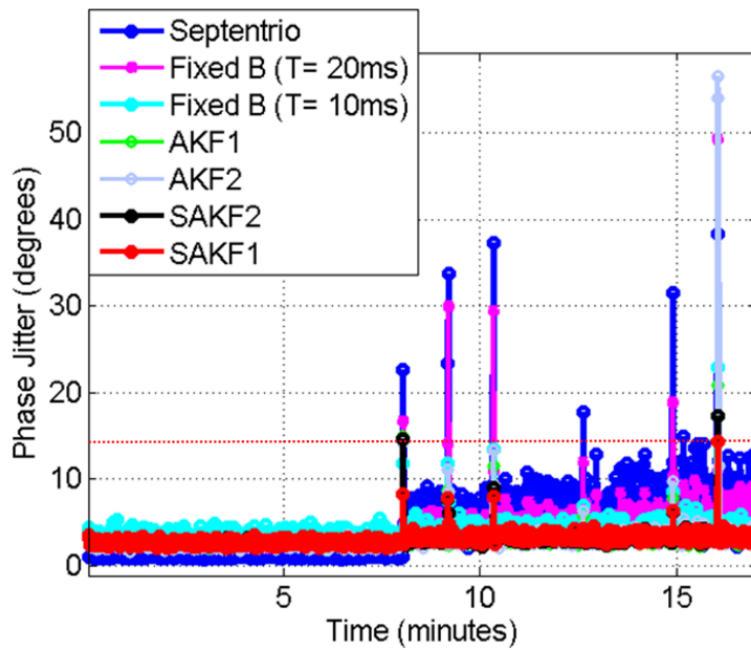


Figure 6-4: Phase jitter comparison (the red line is indicating the 15 degree tracking threshold).

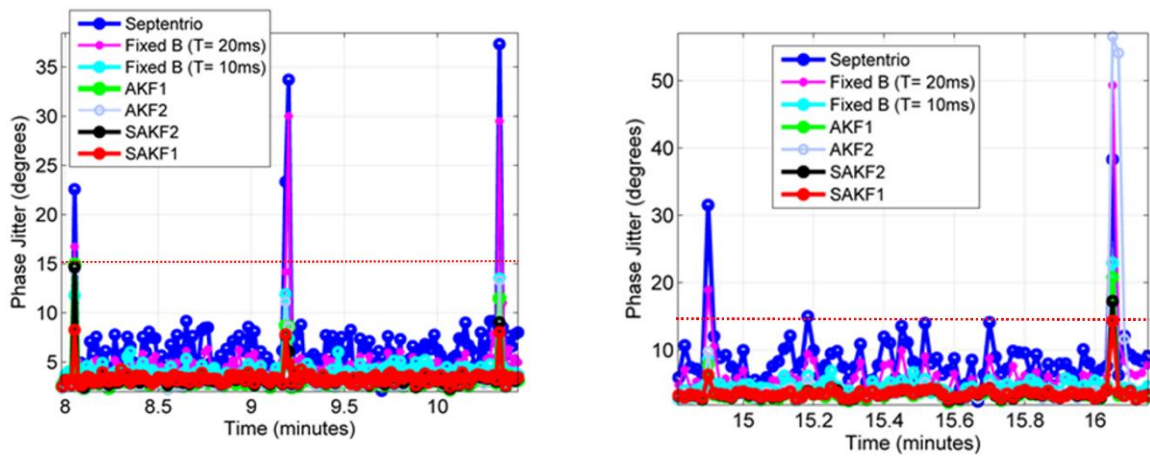


Figure 6-5: (Left) Zoomed versions of the Phase jitter in Figure 6-4 from time equal to 8 to 8.5 minutes (Right) Zoomed versions of the Phase jitter in Figure 6-4 from time equal to 14.5 to 16.5 minutes.

In Figures 6-6 and 6-7 the PLI values and the percentages of PLI samples below 0.86 are shown for the various tracking algorithms.

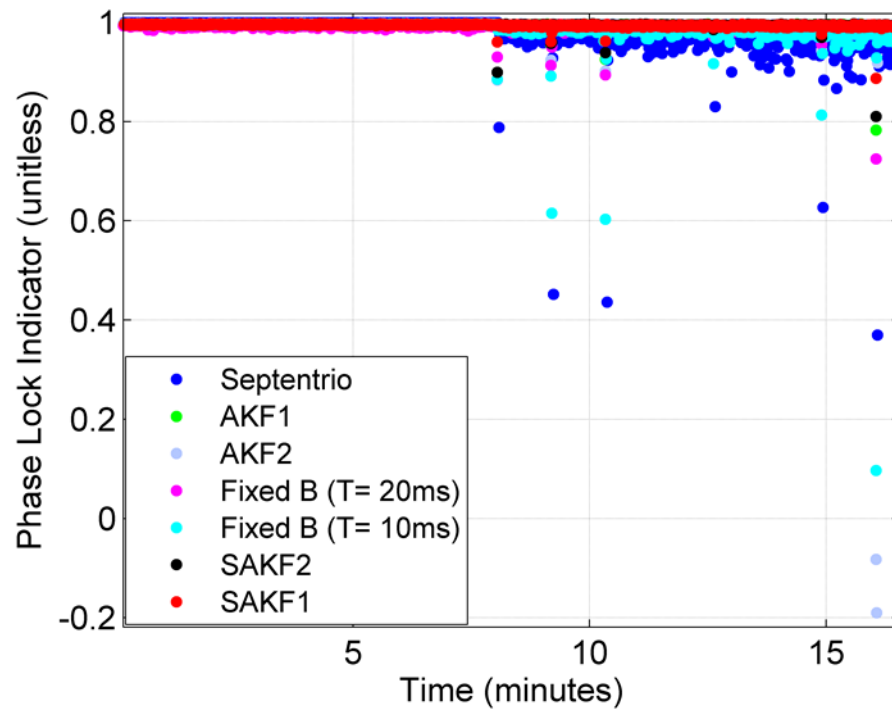


Figure 6-6: PLI comparison.

The improvement given by the SAKFs algorithms is clear from Figure 6-7 where it can be seen that for SAKF1 the percentage of PLI samples below the tracking threshold of 0.86 is equal to zero.

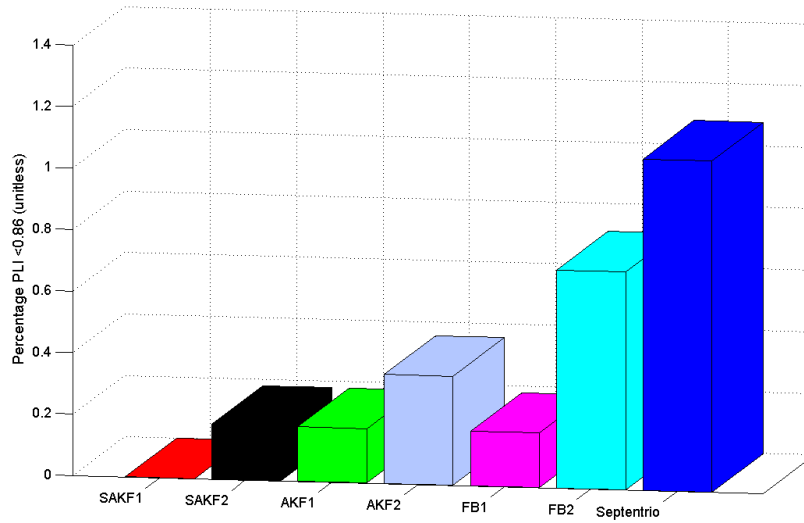


Figure 6-7: PLI comparison (FB1 and FB2 refer to the PLL with fixed bandwidth and T_s equal to 20 and 10 ms, respectively).

In order to assess the algorithms' capabilities in following the signal dynamics, the frequency jitter, computed as the standard deviation of the Doppler frequency over 4 seconds (Kamel, 2010), and the mean frequency jitter are compared in Figures 6-8 (left and right side, respectively) for all the algorithms. The KF based tracking schemes allow reducing the noise present in the carrier Doppler if compared with the fixed bandwidth PLLs. The scheme with fixed bandwidth and time of integration equal to 20 ms shows higher levels of mean Doppler noise.

On the contrary, the SAKF1 PLL achieves the best performance in terms of Doppler noise reduction. It is worth underlining that the performance of SAKF1 and SAKF2 are very close. Therefore SAKF2 represents a good alternative when the high computation cost, required by the continuous scintillation indices computation performed by SAKF1 has to be avoided.

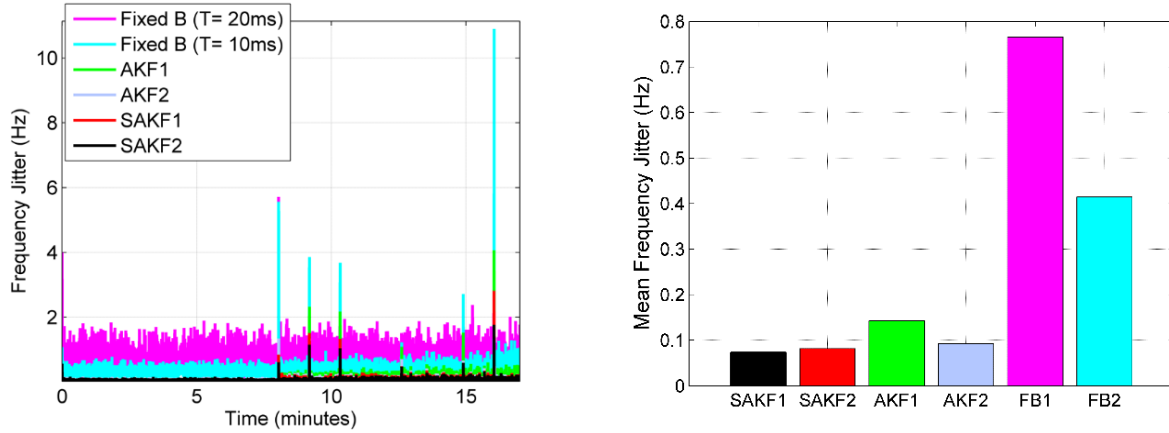


Figure 6-8: (Left) Frequency jitter comparison; (Right) Mean frequency jitter comparison (FB1 and FB2 refer to the PLL with fixed bandwidth and T_c equal to 20 and 10 ms, respectively).

By applying the approach in Xinhua et al. (2015), the steady state equivalent carrier bandwidth for the KF tracking based algorithms can be computed. The bandwidth values for the AKF1, SAKF1 and SAKF2 cases are reported in Figure 6-9. It can be observed that the SAKF algorithms increase their bandwidth when the phase scintillation values increase in order to cope with the dynamics induced by scintillation.

On the contrary, the AKF1 changes its bandwidth only according to the variation of signal intensity.

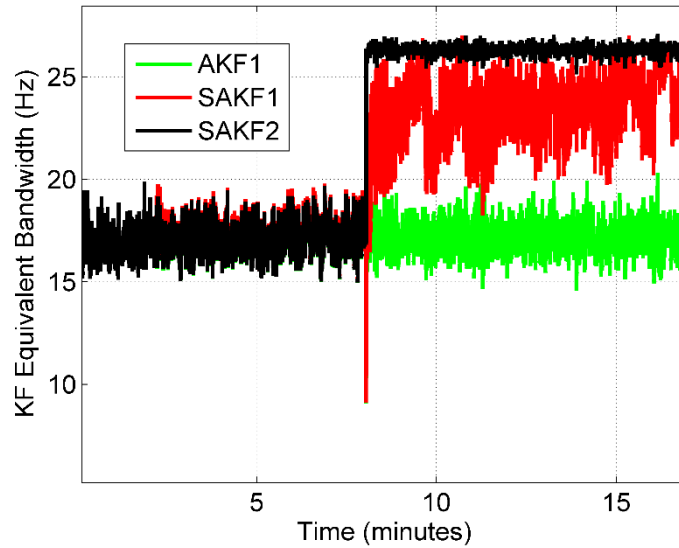


Figure 6-9: Noise equivalent bandwidth comparison.

The bandwidth variation for SAKF2 is quite abrupt since, as explained in Section 5.5, the parameters to tune the KF are selected from a set of four predefined cases. Smoother transitions could be achieved by increasing the number of pre-defined scenarios to tune the KF.

Finally the scintillation monitoring capabilities of the proposed algorithms are assessed by comparing the values of $S4$ and Φ_{60} , estimated as described in Sections 4.2 and 4.3, to the ones provided by the state of art Septentrio receiver, used as benchmark. Figure 6-10 shows the good agreement, for all considered tracking schemes, between the estimated values and the ones used as reference for all KF based tracking algorithms.

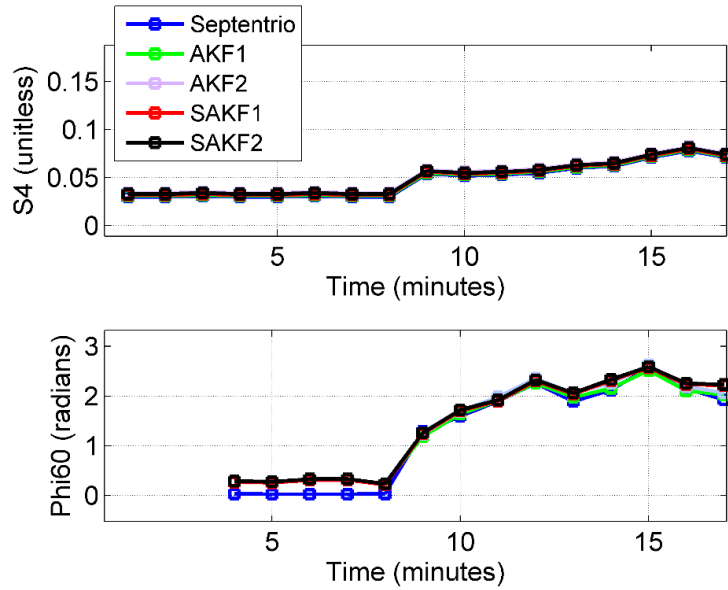


Figure 6-10: (Top) Amplitude scintillation index (S4) comparison; (Bottom) Phase Scintillation index (Phi60).

6.2 Real equatorial scenario: experimental set up

To assess the performance of the proposed algorithms under an equatorial scenario, real data has been used. The latter, provided by the Joint Research Centre of the European Commission, was collected by installing an USRP N200 front-end in Hanoi, Vietnam (lat 21° 2' 0" N / long 105° 51' 0" E) during a period of intense solar activity. This front end is an older version of the one described in the previous section. Also this USRP includes a Temperature-Controlled Crystal Oscillator (TCXO), which, as explained in the previous section, is not optimal for scintillation monitoring. Consequently, also this time an external oscillator has been necessary and a 10MHz Rubidium oscillator was used to drive the front-end. The data was collected at 5 MSamples/sec in the L1 band each day after sunset local time in order to monitor the peak of the ionospheric disturbances.

A Septentrio PolaRx4 receiver was colocated with the USRP to store additional data such as C/N₀, azimuth and elevation of the satellites in view. However the Septentrio PolaRx4

receiver is not an ISMR, so it does not provide scintillation indices. For this reason, a replay process of the USRP logged data has been performed to obtain, through a Septentrio PolaRxS ISMR, the scintillation indices and 50 Hz In phase (*I*) and in Quadrature (*Q*) samples. The scheme of the experimental set up is reported in Figure 6-11 while the data collection details are summarized in Table 6-3.

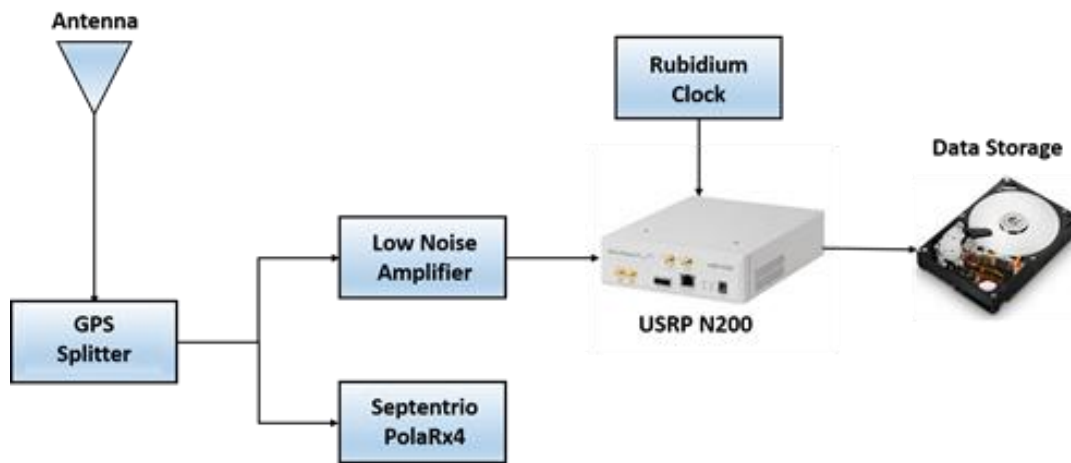


Figure 6-11: Experimental set up scheme.

Table 6-3: Data collection details.

Parameter	Value
Signal	GPS L1
IF	0
Sample Type	Complex
Bits per sample	16
Sampling Rate	5 MHz

6.2.1 Data processing methodology

A data set affected by strong scintillation, collected on April 16th 2013 has been selected for the algorithms assessment. The sky-plot for this data set is shown in Figure 6-12. For the selected data-set, 10 satellites are available, of which 6, indicated in Figure 6-12 with red circles, are affected by scintillation.

To assess the algorithms, four satellite links with different levels of scintillation have been selected. They are **SV1**, **SV7**, **SV8**, and **SV28**. The tracking algorithms used to process the data are summarized in Table 6-4. Also in this case AKF1 is a classical adaptive KF tracking with **Q** a priori fixed, as in Macchi (2010), and with the elements of **R** obtained as detailed in Section 5.3. In this case AKF2 has the same characteristics as AKF1 but without applying the scaling factor, introduced in equation 5.34, to **R**. In this case AKF2 has been included to assist in better analysing the single contributions given by the covariance matrix tuning and the measurement noise tuning to the performance improvement. For all KF based tracking schemes a time of integration T_c , equal to 20 ms, has been set. A time of integration higher than the high latitude scenario has been selected considering that under the equatorial scenario the dynamics are generally more moderate. The setting of the tracking loops with fixed bandwidth has been selected following the same criteria explained in Section 6.1.1.

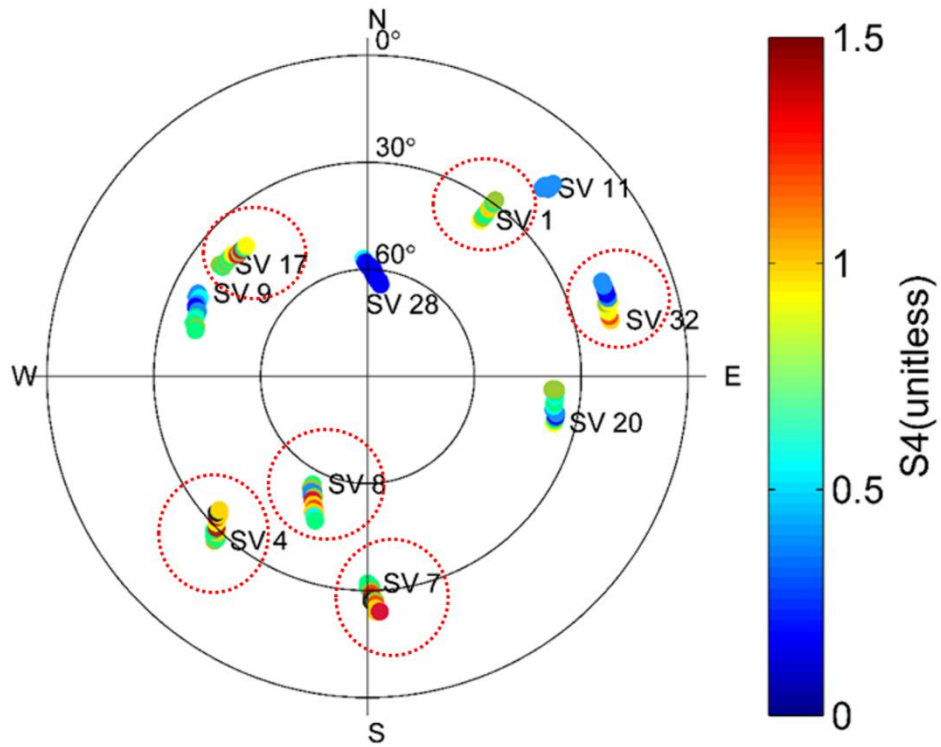


Figure 6-12: Skyplot for the data set collected in Hanoi (Vietnam) between 13.20-13.40 UTC on April 16th 2013.

Table 6-4: Tracking architecture details.

Tracking Architecture	Q	R	T_c (ms)	B (Hz)
AKF1	A priori fixed	Variable (eq 5.26 and 5.27) without scaling factor	20	Not applicable
AKF2	A priori fixed	Variable with scaling factor (eq. 5.34)	20	Not applicable
SAKF1	Variable	Variable with scaling factor (eq. 5.34)	20	Not applicable
SAKF2	Variable	Variable with scaling factor (eq. 5.34)	20	Not applicable
Fixed B (T=20ms)	Not applicable	Not applicable	20	4
Fixed B (T=10ms)	Not applicable	Not applicable	10	15

6.2.2 Test results

As mentioned in Section 6.2.1, four satellites, namely **SV1**, **SV7**, **SV8** and **SV28**, have been selected for the algorithms' assessment. As shown in Figure 6-13 and Figure 6-14, where the satellites' scintillation indices are reported, SV1 is affected by moderate/strong amplitude scintillation and weak phase scintillation while SV7 and SV8 are affected by both very strong amplitude and phase scintillation. Finally SV28, apart from the first two minutes, is almost scintillation free. This satellite has been selected to observe the algorithm performance also under quiet conditions. For SV7, namely the satellite link affected by the most severe scintillation level, no *Phi60* values are provided by the Septentrio PolaRxS receiver from minute 6 to minute 11. This is due to temporary losses of lock, which induce the phase detrending filter to reset, producing, as a consequence, gaps in the output *Phi60*. Indeed, as previously mentioned, it is necessary up to four minutes for this filter to converge. Consequently, to ensure reliable and continuous scintillation monitoring capabilities the losses of lock and cycle slips should be minimized.

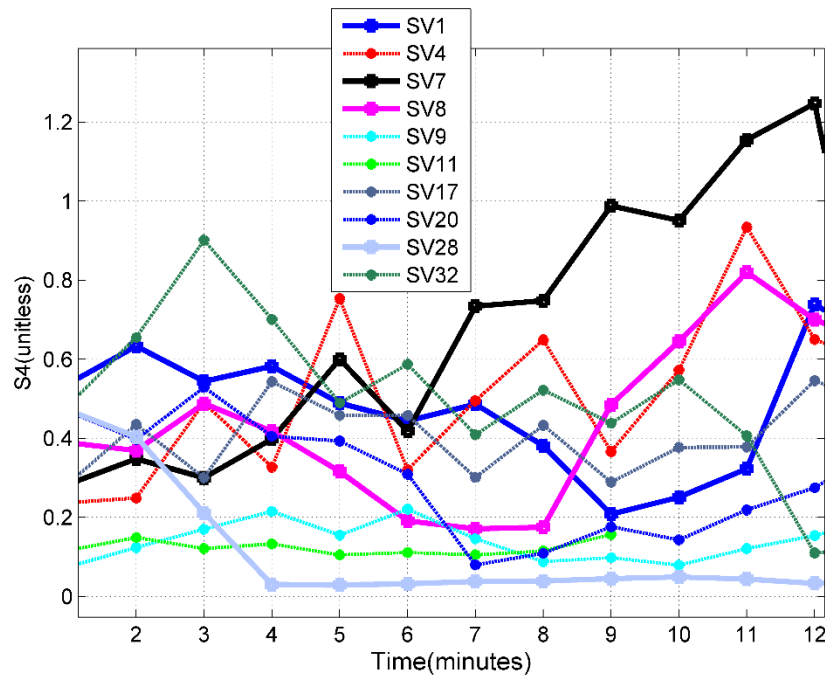


Figure 6-13: Time series of the amplitude scintillation (S4) index for the GPS L1 data set collected in Hanoi (Vietnam) on April 16th 2013.

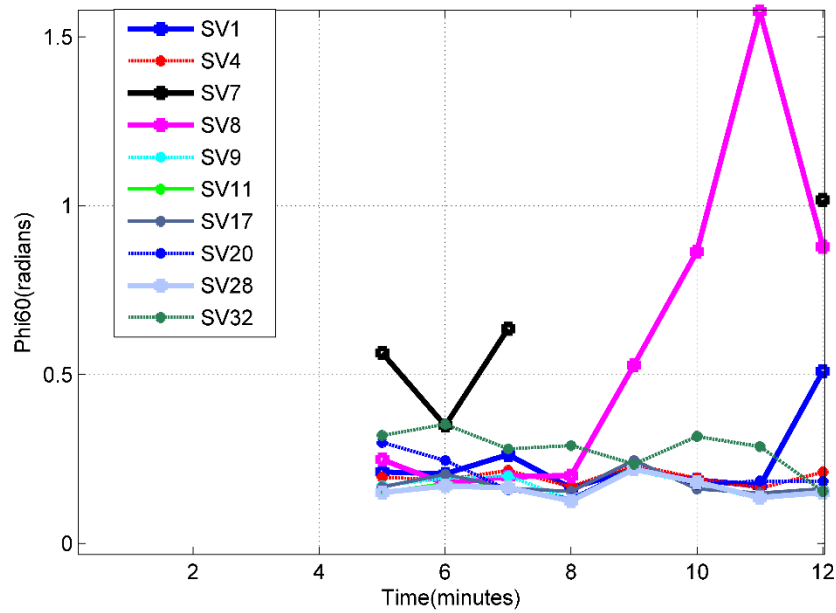


Figure 6-14: Time series of the phase scintillation index (Phi60) for the GPS L1 data set collected in Hanoi (Vietnam) on April 16th 2013.

To compare the tracking performance of the above algorithms, the phase jitter for SV7 was selected because this satellite was affected by the most severe level of scintillation. In fact it was the only link where a loss of lock occurred for the Septentrio receiver. The computed result is shown in Figure 6-15. Also in this case, the phase jitter, namely the standard deviation of the phase discriminator output, has been computed over temporal windows of 4 seconds. As it can be seen in Figure 6-15, SAKF1 and SAKF2 show the best performance in terms of phase jitter reduction. Moreover the two algorithms achieve very close performance. On the other hand, AKF1 shows poor performance and indeed, after the 7th minute, the AKF1 model fails, leading to various losses of lock occurring when the phase jitter is over 15 degrees.

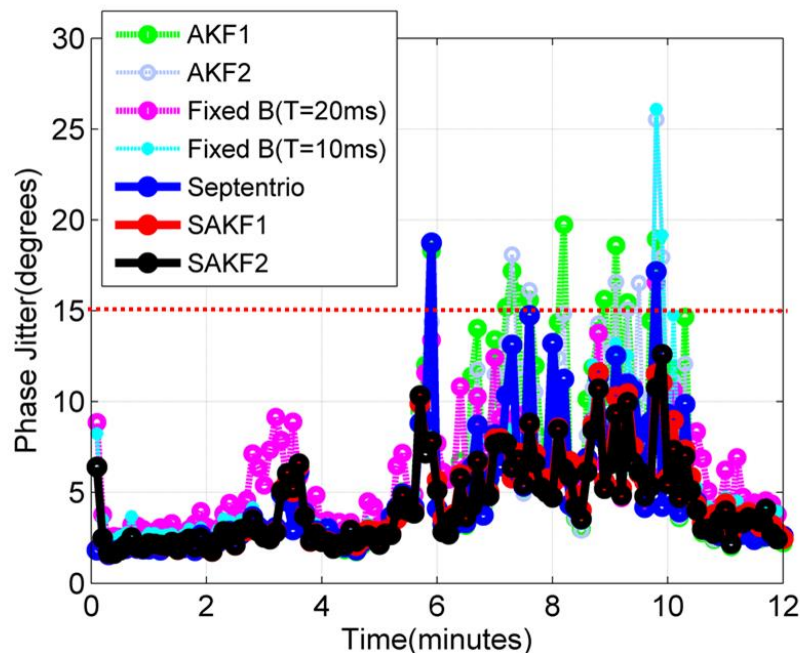


Figure 6-15: Phase Jitter comparison for the satellite link SV7 characterized by both strong phase and amplitude scintillation. The red line indicates the 15 degree tracking threshold.

Due to bad modelling, the AKF1 shows even worst performance than the traditional tracking with fixed bandwidth and $T_c = 20\text{ms}$. Finally, SAKF1 and SAKF2 outperform also the Septentrio PolaRxS ISMR. In Figure 6-16 Φ_{i60} and the strength of the phase scintillation spectrum at 1 Hz (T) for SV7 (computed by the SAKF1 to adjust the KF covariance matrix) are shown. It is interesting to observe the clear correlation between the two parameters over time. Both parameters, as well as the slope of the phase scintillation spectrum p , have been computed by using sliding windows integrated at each time of integration (T_c). T has been estimated by evaluating the PSD of the detrended accumulated carrier phase FFT at 1 Hz. The slope of the spectrum p has been estimated by computing the slope of a straight line obtained by a linear fit to the detrended carrier phase PSD (Aquino et al., 2007).

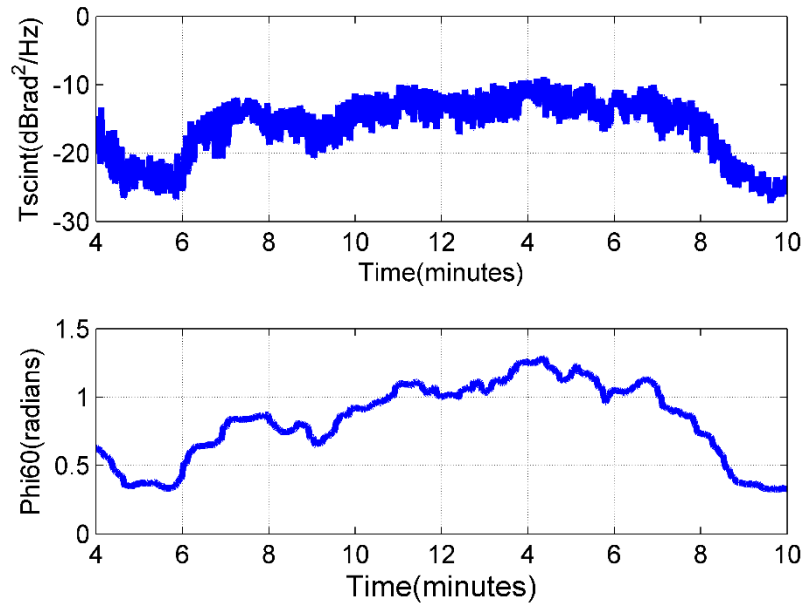


Figure 6-16: (Top) Spectral strength of the phase noise at 1 Hz (T) and (Bottom) Phase scintillation index (Φ_{i60}) for SV7.

In Figures 6-17 and 6-18 the PLI values and the percentages of PLI samples below 0.86 are shown for the various tracking algorithms, for SV1, SV7, SV8 and SV28. SAKF1 and SAKF2 achieve the best performance for the satellites affected by both strong phase and amplitude scintillation, namely for SV7 and for SV8. In the case of SV1, where the amplitude scintillation is dominant, the performances of the KF based algorithms and also of the Septentrio receiver are very close. To better quantify the advantage of using the proposed algorithms, Table 6-5 demonstrates the improvement percentages (in terms of reduction in the occurrence of PLI samples below 0.86), achieved by the SAKF2 algorithm with respect to the other tracking schemes. The SAKF1 algorithm has not been included in the comparison since, as shown in Figure 6-18, the performances of SAKF1 and SAKF2 are very close. It is clear that the SAKF2 outperforms all the other tracking schemes.

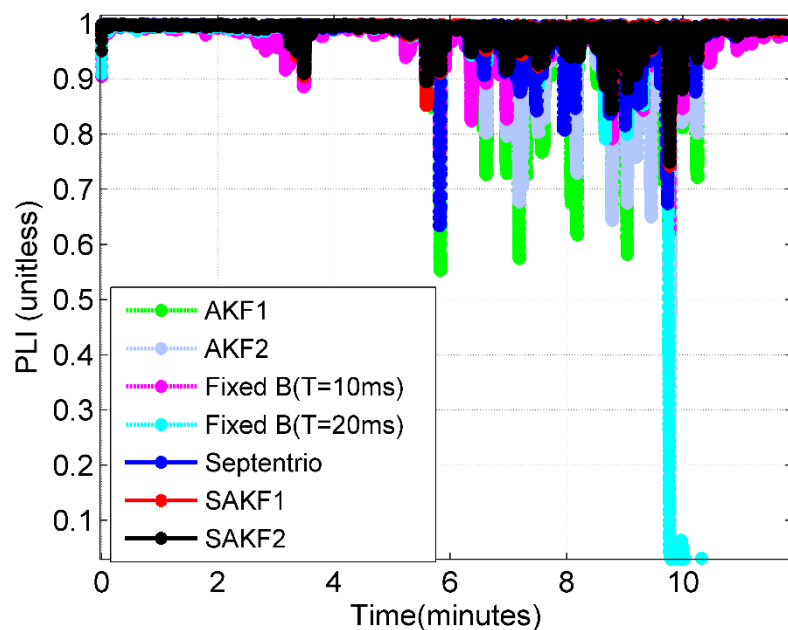


Figure 6-17: Phase Lock Indicator (PLI) comparison.

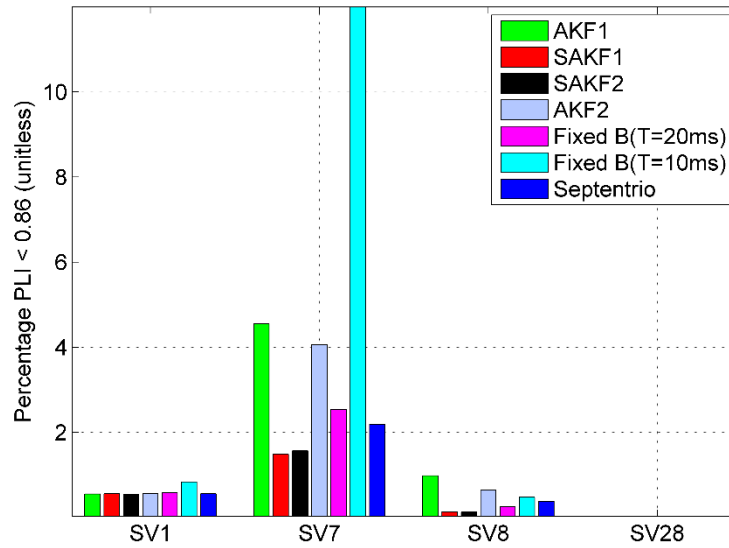


Figure 6-18: Percentage of PLI samples below 0.86.

Table 6-5: Improvement in the occurrence of PLI values below 0.86.

Improvement (%)	SAKF2 vs commercial ISMR	SAKF2 vs AKF1	SAKF2 vs Fixed Bandwidth PLL ($T_c=20\text{ms}$)	SAKF2 vs Fixed Bandwidth PLL ($T_c=10\text{ms}$)
SV1	0.5	0.2	4.7	33.3
SV7	32.1	67	41.4	92
SV8	64.7	86	45.4	71
SV28	0	0	0	0

To compare the tracking performance in terms of agility in following the signal dynamics, in Figure 6-19 the Doppler shift obtained by the different algorithms is shown. All KF based tracking schemes outperform the classical fixed bandwidth PLL/DLL in terms of Doppler noise reduction. Moreover it can be observed at about minute 10 a loss of lock occurs for the scheme with fixed bandwidth and $T_c=10\text{ms}$. In Figure 6-20 the mean values of the Doppler shift standard deviations, computed over the observation period, are reported for SV1, SV7, SV8 and SV28. In this case all KF based tracking schemes achieve close values, outperforming the algorithms with fixed bandwidth.

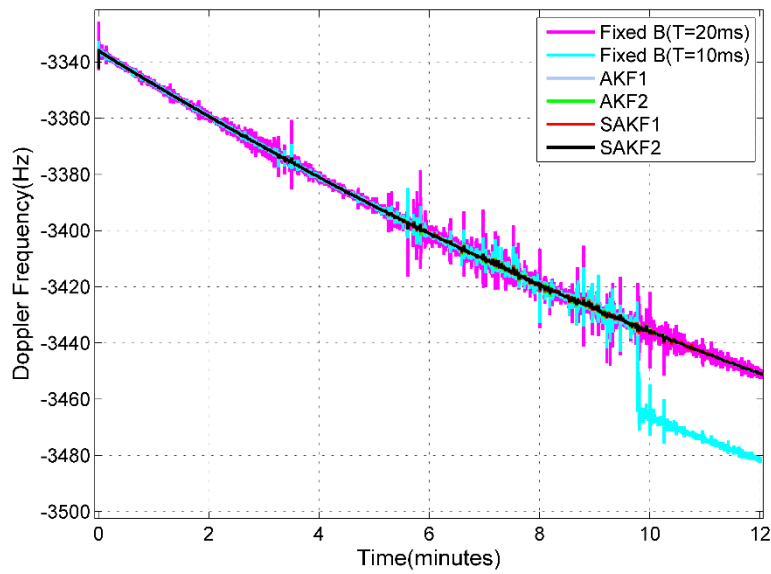


Figure 6-19: Doppler Frequency comparison.

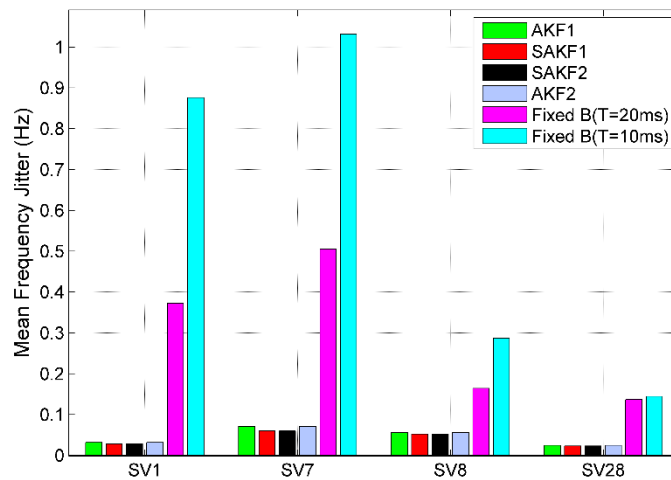


Figure 6-20: Mean frequency jitter over the interval of observation.

The equivalent carrier bandwidth for SV7 (the satellite link affected by the strongest level of scintillation) is reported in Figure 6-21 for AKF1, SAKF1, SAKF2. Again it is worth reminding that 4 minutes are necessary for the SAKF1 and SAKF2 to start computing the

parameters necessary to tune their dynamic models. After the fourth minute the effect of the dynamic model adjustment is reflected in the increase of the bandwidth values with respect to the values of AKF1. Also in this case, for SAKF2 the variations in the bandwidth values are less marked due to the use of the four pre-defined dynamic models. Both SAKF1 and SAKF2 increase their equivalent bandwidth to achieve a higher agility in following the dynamics when the phase variation is stronger, thanks to the fact that the covariance matrix includes the phase scintillation contribution. At the same time, the deep fading in the bandwidth values show also the good response of SAKF1 and SAKF2 to the C/N_0 variations. On the contrary, the AKF1 bandwidth values are much lower due to the fact that it is only adjusted according to the C/N_0 variations.

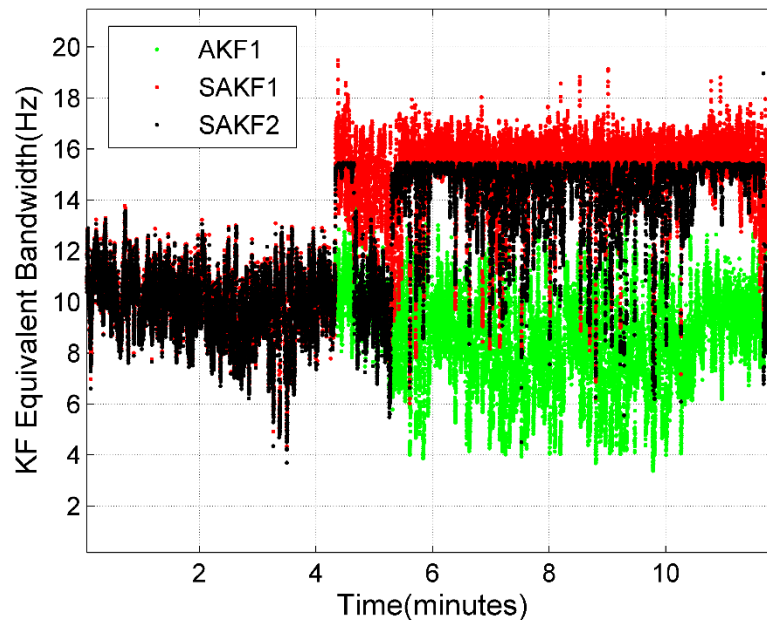


Figure 6-21: Equivalent bandwidth values comparison for SV7.

These lower values do not allow following the signal dynamics, producing a poorer tracking performance in the presence of strong phase scintillation. Furthermore, the correlator outputs and the accumulated phase obtained by the proposed tracking schemes have been used to compute the scintillation indices $S4$ and Φ_{60} as in Van Dierendonck (1993). As an example, the scintillation indices computed for SV7 are shown in Figure 6-22 along with their counterparts provided by the commercial ISMR used as benchmark. The $S4$ values obtained by the KF tracking schemes are in good agreement with the values provided by the commercial ISMR.

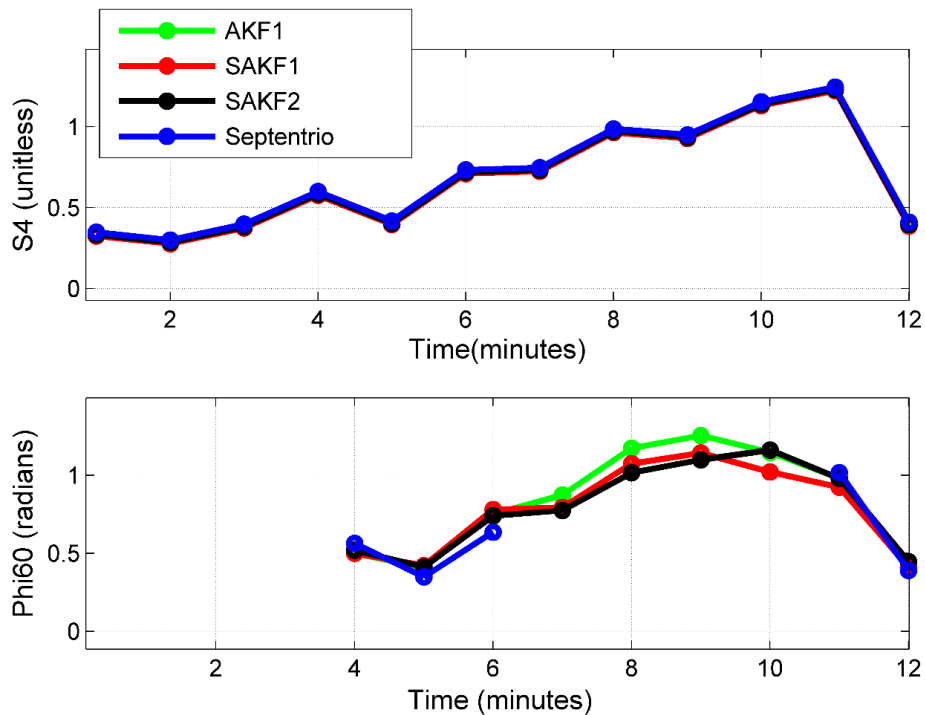


Figure 6-22: (Top) Amplitude scintillation indices and (Bottom) Phase scintillation indices for SV7.
Due to temporary losses of lock, data gaps occur in the Φ_{60} time series provided by the Septentrio receiver.

As previously mentioned, the Septentrio receiver shows data gaps in the time series of Φ_{60} , due to losses of lock. Figure 6-22 (lower part) shows the comparison for Φ_{60}

between the commercial ISMR and the implemented tracking schemes. For the implemented tracking schemes, the Φ_{60} is actually computed by a 1 minute sliding window at every integration time (T_c). The Φ_{60} values for AKF1, SAKF1 and SAKF2 shown are actually an average of the Φ_{60} values computed during one minute, and presented at the end of each minute. As shown in Figure 6-22, in this way it is possible to obtain continuous information of Φ_{60} not only for SAKF1 and SAKF2, which present higher tracking performance, but also for AKF1, which, on the contrary, experiences several losses of lock. Therefore, although for AKF1 more losses of lock occurred than for the ISMR, by computing Φ_{60} using sliding windows, the scintillation monitoring capabilities of the algorithms are not affected and no Φ_{60} samples are missed.

6.3 Equatorial simulated scintillation scenario

A further test under equatorial scintillation has been performed in order to investigate the effectiveness of using a low cost TCXO based receiver, namely the **Connected Autonomous Space Environment Sensor (CASES)** receiver (O'Hanlon et al, 2011), for scintillation monitoring purposes.

The CASES receiver is a complete box receiver based on a real time GPS software defined receiver designed for ionospheric scintillation monitoring by the Cornell University, the University of Texas, and ASTRA to offer a low cost, flexible and low power consuming instrument for ionospheric studies. It is worth pointing out that it was not possible to have access to the source code of the CASES but only to its output, as for a standard hardware receiver.

The main purpose of this test was to assess the performance of the CASES receiver as compared to a high grade commercial scintillation monitoring receiver, namely the Septentrio PolaRxS. This was achieved by comparing the scintillation indices provided by the two receivers under the same scintillation scenario.

The CASES receiver has a temperature-compensated crystal oscillator (TCXO) as internal clock, rather than a more accurate oven controlled crystal oscillator (OCXO) which, as previously underlined, is usually adopted by high grade (and more costly) commercial scintillation monitoring receivers, such as the Septentrio PolaRxS.

Therefore, in order to remove the clock noise effect, the CASES includes an algorithm, as detailed in Section 4.3, to estimate the clock phase noise contribution by using a scintillation-free link as reference channel. Specifically, it removes the phase error contribution due to the clock, and isolates the scintillation contribution, by differencing the carrier phase time series of a scintillation-free link with the ones obtained for the link affected by scintillation.

Furthermore, the same algorithm exploited by the CASES receiver has been implemented also into the software receiver based algorithms proposed by this thesis in order to provide scintillation monitoring capabilities even when only a low cost clock is available.

Finally, the test presented herein allowed to further assess and compare the tracking performance of the proposed tracking algorithms with the traditional tracking schemes, the CASES and the Septentrio PolaRxS receivers.

6.3.1 Experimental test set-up

In order to test the CASES receiver for the high latitude scintillation case detailed in Section 6.1, the Spirent GSS8000 GNSS simulator was used to simulate a GPS L1 signal. The simulator was simultaneously connected, through a GPS splitter, to the CASES receiver, to a Septentrio PolaRxS receiver, used as benchmark, and to the Universal Software Radio Peripheral (USRP) N210 front end whose characteristics have been described in Section 6.1. A general overview of the experimental set-up is shown in Figure 6-23.

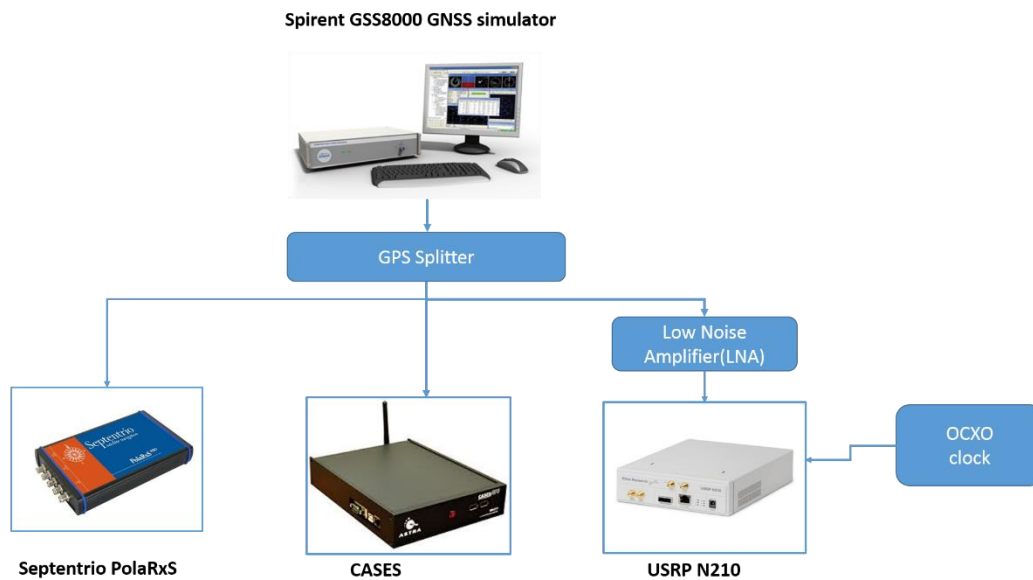


Figure 6-23: Experimental set-up scheme.

The CASES receiver functionalities can be controlled through specific configuration files. The Single Board Computer (SBC) configuration file allows controlling the high level functionalities of the receiver (i.e. scintillation parameters, number of satellites to track,

etc.), while the Digital Signal Processor (DSP) configuration file allows controlling the receiver low-level functionalities (i.e. tracking parameters, elevation mask, etc.). The CASES receiver DSP configuration file has been modified so as to set the same Phase Locked Loop (PLL) bandwidth value ($B = 15\text{Hz}$) and time of integration ($T_c = 10\text{ ms}$) used in the PolaRxS receiver by default. As underlined above, the main goal was to assess the scintillation monitoring capabilities of the CASES by using the values obtained by the Septentrio PolaRxS receiver as a reference.

For this purpose the Spirent simulator has been used in conjunction with real equatorial data. In fact, as explained in Section 6.1, the Spirent simulator can receive as input a file, namely the User Command Defined file (.ucd), containing amplitude and phase variations to be superposed to the nominal signal in order to generate signals affected by scintillation. To simulate a realistic scenario, amplitude and phase variations under scintillation were obtained from real field data. Specifically, they have been extracted from high rate (50 Hz) data collected by a Septentrio PolaRxS receiver installed in Presidente Prudente ($22.1^\circ\text{S}, 51.4^\circ\text{W}$), Brazil, during a period of intense scintillation activity on 24th November 2011. The high rate In-phase (I) and Quadrature (Q) components along with the carrier phase have been detrended in order to remove any source of noise apart from scintillation and to extract the amplitude and phase fluctuations produced by scintillation. The detrending operation was the same that is usually performed for the computation of the *Phi60* index, as described in Section 4.3. Therefore, the phase variations have then been used to format the .ucd file so that one of the available satellites was affected by scintillation.

6.3.2 Test results

The scintillation indices recorded respectively by the CASES and the Septentrio receivers for the simulated satellite link affected by scintillation are shown in Figure 6.24.

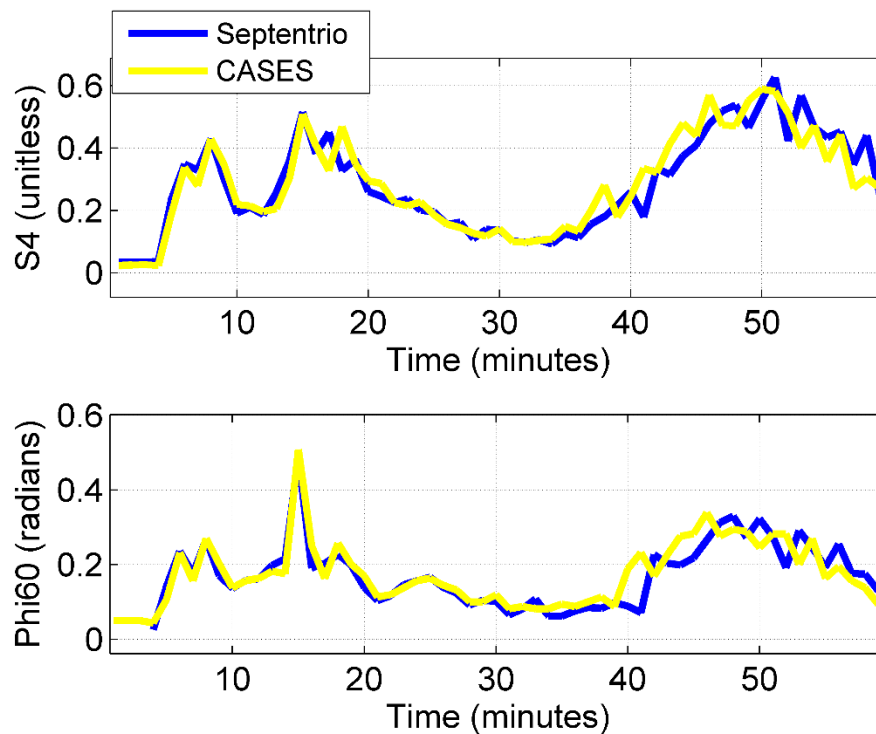


Figure 6-24: Scintillation indices recorded by the Septentrio and CASES receivers for the simulated scenario.

It can be seen that the scintillation indices provided by the CASES receiver are very close to those given by the Septentrio receiver. Therefore, despite its low quality clock, the CASES provides good scintillation monitoring capabilities, comparable to the high grade scintillation monitoring receivers, namely the Septentrio PolaRxS.

Therefore to evaluate the tracking performance, the first 15 minutes of data have been considered. The IF data have been processed by the software receiver configured with the tracking algorithms detailed in Table 6-6.

Table 6-6: Tracking algorithms.

Tracking Architecture	Q	R	T_c (ms)	B (Hz)
AKF1	A priori fixed	Variable with scaling factor (eq. 5.34)	20	Not applicable
SAKF1	Variable	Variable with scaling factor (eq. 5.34)	20	Not applicable
SAKF2	Variable	Variable with scaling factor (eq. 5.34)	20	Not applicable
Fixed B ($T_c= 20\text{ms}$)	Not applicable	Not applicable	20	4
Fixed B ($T_c= 10\text{ms}$)	Not applicable	Not applicable	10	15

As for the tests described in Sections 6-1 and 6-2, the phase jitter has been evaluated and compared for the different algorithms/receivers. It is also worth pointing out that the CASES, as the Septentrio, provides 50 Hz I and Q correlator values. Therefore also for the CASES it has been possible to evaluate the phase jitter and the PLI. In Figure 6-25, where the phase jitter values are compared, it can be noticed that again the SAKFs algorithms achieve the best performance in terms of noise reduction, maintaining the phase jitter below the 15 degrees. This is more evident in Figure 6-26, where the corresponding phase/amplitude scintillation peak of the phase jitter plot of Figure 6-25 is reported in detail. On the other hand the CASES achieves the worst performance. However it should be taken into account that the higher clock noise, due to the lower clock quality, could be responsible for this degradation in the tracking performance.

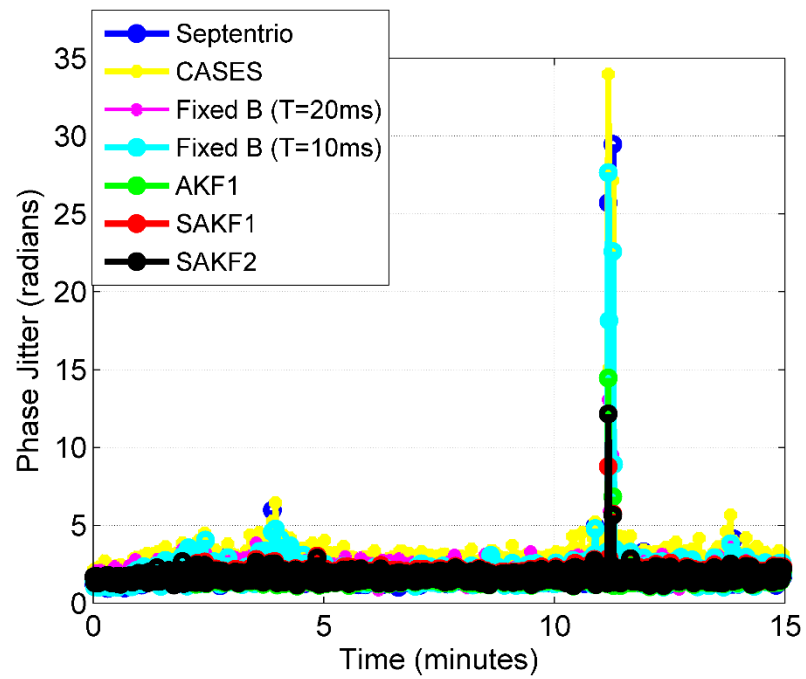


Figure 6-25: Phase Jitter comparison.

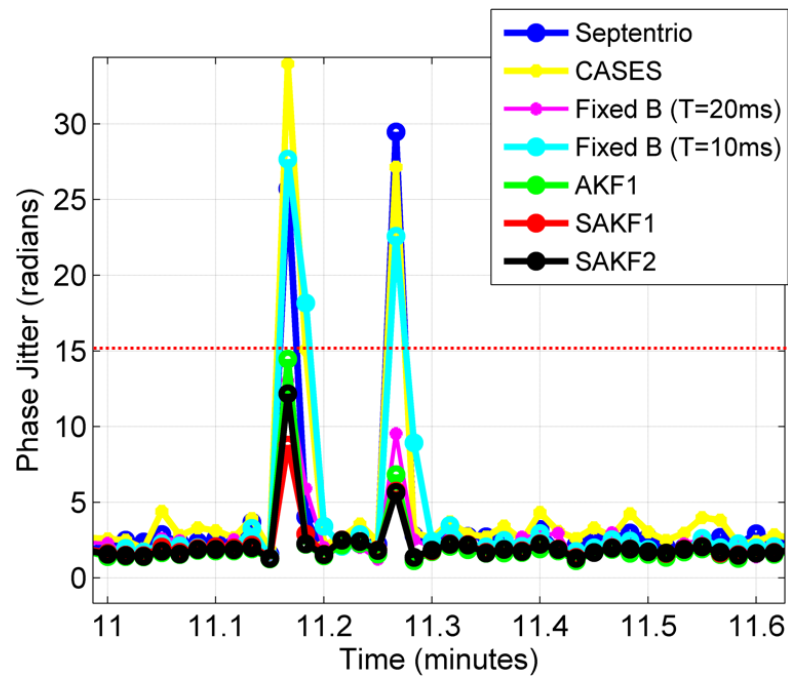


Figure 6-26: Zoomed version of the Phase Jitter in Figure 6.25.

Figure 6-27 reports the PLI values estimated for the analyzed tracking schemes/receivers and the percentage of PLI values below 0.86 degrees. These plots clearly show the improvement produced by the SAKF in terms of tracking robustness.

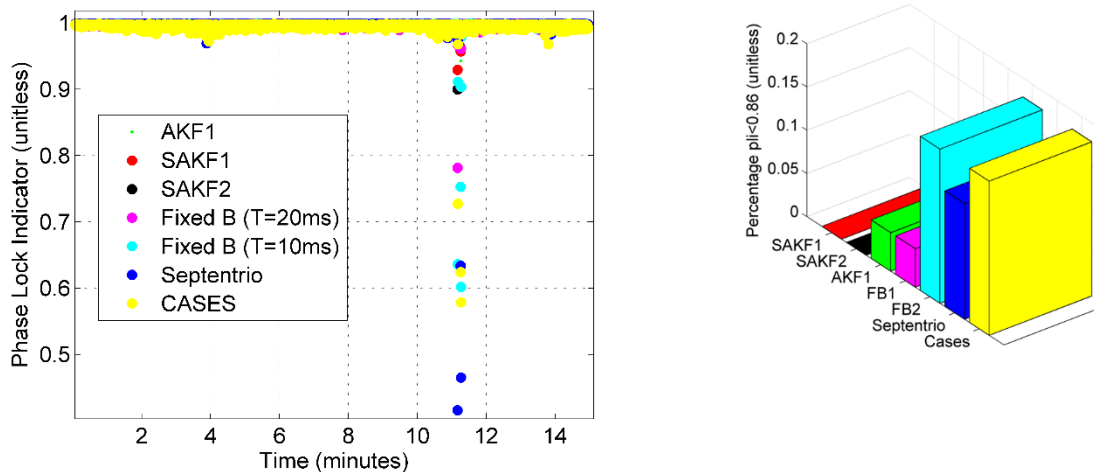


Figure 6-27: (Left) Phase Lock Indicator (PLI) comparison; (Right) Percentage of PLI sample below 0.86 degrees.

In order to analyze the ability in tracking the high dynamics produced by scintillation, the Doppler Frequency is compared and reported in Figure 6.28. The scheme with fixed bandwidth and time of integration equal to 20 ms shows the noisier Doppler output, while again the SAKFs scheme achieves the best performance, as clearly seen from Figure 6-29, where the Doppler jitter (left side) and the mean Doppler jitter (right side) are shown.

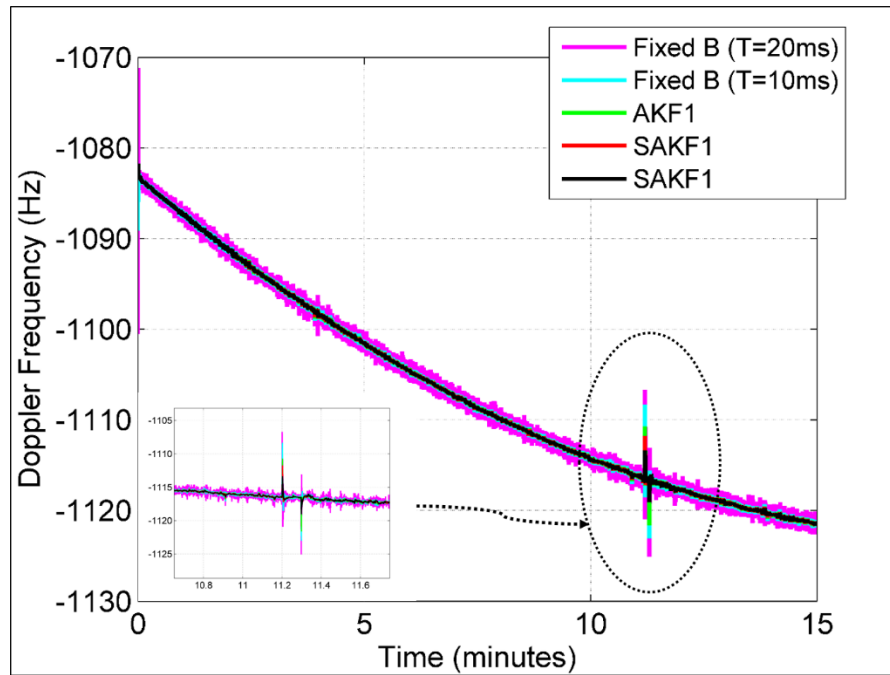


Figure 6-28: Doppler Frequency comparison.

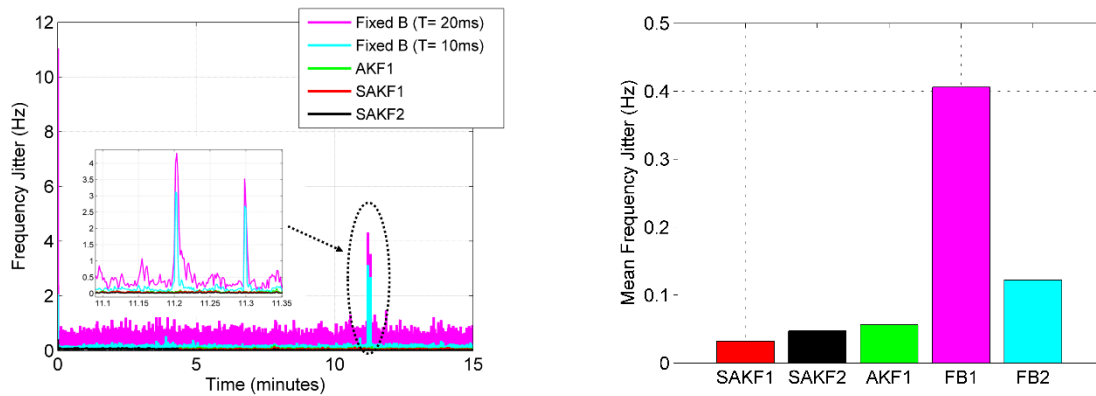


Figure 6-29: (Left) Frequency Jitter comparison and (Right) Mean Frequency Jitter.

Therefore, also this experiment showed the advantage of optimizing the KF tuning in presence of scintillation. In fact, also in this case SAKF1 and SAKF2 show an improved robustness if compared with traditional tracking schemes, as well as with the PolaRxS

Septentrio receiver. Finally, it has been shown that despite its low quality clock, the CASES receiver achieves good tracking and scintillation monitoring performance.

6.4 Summary

This chapter discussed the results obtained by assessing the tracking algorithms proposed in Chapter 5, SAKF1 and SAKF2, under high and equatorial latitudes scenarios. It was shown that the above algorithms allow achieving higher performance than traditional tracking schemes by taking advantage of the knowledge about the scintillation scenario obtained through the continuous monitoring of a number of scintillation indicators. Moreover it has been noticed that a main issue of commercial ISMRs is the presence of data gaps in the phase scintillation index time series when losses of lock occur. The reason is that because of the loss of lock, the phase detrending filter, used for the index computation, has to be reset, inducing data gaps in the estimation of these parameters. To avoid this issue a technique has been proposed which uses a sliding window for the index computation, allowing continuous phase indices information when commercial ISMRs fail. Finally, the performance of a TCXO based receiver has been assessed under scintillation. The receiver, through a specific algorithm, can reduce the clock noise, achieving, for the case study examined herein, good scintillation monitoring capabilities despite its internal low cost TCXO.

Chapter Seven: Conclusions and Recommendations

This thesis involves research aiming to investigate the design of novel carrier algorithms for GNSS tracking under ionospheric scintillation.

Scintillation is one of the most significant ionospheric related phenomena, which can impair GNSS positioning by degrading signal tracking and even causing complete loss of signal lock. High accuracy positioning can be particularly badly affected, and regions of low and high latitudes are the most susceptible to the phenomenon.

The tracking stage is the receiver part most vulnerable to deep signal fading and fast Doppler variations due to scintillation. This can be a main issue especially for Ionospheric Scintillation Monitoring Receivers (ISMRs) which are usually placed in locations where scintillation activity can be very intense and challenging for the receiver tracking. Consequently, under such harsh scenarios poor tracking performance will be translated into poor scintillation monitoring capabilities.

This research work addressed the design of tracking algorithms optimized to be robust under ionospheric scintillation and especially designed for ISMRs which, by definition, should provide reliable and continuous information about the ionosphere.

In particular the work focused on the smart tuning of Kalman Filter (KF) based tracking algorithms so as to increase their robustness under both amplitude and phase fluctuations associated with scintillation.

The present chapter summarizes the main outcomes of this research work and provides recommendations for future work that may lead to further enhancing the algorithms developed and proposed by this thesis.

7.1 Conclusions

On the basis of the investigation and outcomes of this research work the following conclusions can be drawn:

1. In presence of scintillation traditional carrier phase techniques impose a trade-off between noise reduction and agility in the tracking of high dynamics. A narrow bandwidth (or a longer prediction integration time) is required to reject the noise produced by scintillation fading, while a wider bandwidth (or a shorter prediction integration time) is required to follow the fast signal dynamics induced by phase scintillation. To solve the above trade off, advanced carrier tracking architectures should be exploited.
2. The literature review of tracking algorithms robust under scintillation has revealed that, although a number of research works exploit KF based tracking architectures to cope with the variable conditions imposed by scintillation, little work has been done to specifically optimize the KF algorithms for scintillation scenarios. In fact it has revealed that a priori fixed dynamic model KF tracking algorithms can fail in presence of strong phase and amplitude fluctuations. Therefore, especially when

low noise clocks are used, as in the case of the most common ISMRs, the scintillation noise contribution should not be neglected in the KF modelling.

3. Two novel Scintillation based Adaptive Kalman Filters, indicated as SAKF1 and SAKF2, have been proposed in Chapter 5. They self-tune their covariance matrices according to the detected level of scintillation and self-adapt their measurement noise model to cope with simultaneous phase and amplitude variations. SAKF1 requires a continuous computation of phase scintillation spectral parameters, whereas SAKF2 selects the dynamic model for the specific scenario from a set of a priori defined options.
4. In Chapter 6 it has been shown that SAKF2 allows achieving performance almost comparable to SAKF1 while reducing the computational cost. Both algorithms outperform the classical adaptive KF, traditional PLL/DLL tracking algorithms with fixed bandwidth and even a commercial ISMR when severe amplitude and phase variations occur simultaneously.
5. It has been observed that data gaps can occur in the phase scintillation indices time series provided by commercial ISMRs when temporary losses of lock occur. These gaps are produced by the resetting of the phase detrending filter, used to estimate these parameters. It has also been shown that by computing phase scintillation indices using sliding windows, as done by SAKF1 and SAKF2, it is possible to get

a higher number of samples reducing the probability of having data gaps in the Phi60 computation when losses of lock are present. This approach allows achieving scintillation monitoring performance capabilities higher than with the commercial ISMR used as benchmark but, on the other hand, it increases the required computational cost.

6. The algorithms proposed by this thesis are based on the computation of a number of scintillation indices. If the estimation of these parameters is not reliable because of hardware features, such as low quality clocks, specific algorithms, as discussed in Chapter 6, should be applied in order to minimize the extra noise which is not produced by scintillation.

7.2 Recommendations

In light of the developments brought about by this research work and building on the above conclusions, the following recommendations are suggested for future work

1. Although the algorithms have been used to track GPS L1 signals, the same methodology could be also extended to multi-frequency tracking schemes including those based on a vector tracking architecture. Indeed KFs are usually at the core

of these more complex architectures, which are expected to gain increasing popularity and diffusion with the full development of new constellations. In this regard it should be pointed out that the algorithms proposed by this thesis and published in (Susi et al 2014a) have been already applied to multi-frequency KF based tracking algorithms by Ashwitha et al. (2015). Their work shows that the use of multi-frequency architectures provides only a very marginal improvement to the one already given by the scintillation based adaptive nature of the scheme. However further investigations in this direction are suggested.

2. The algorithms proposed herein have been assessed with real equatorial data and simulated high latitude data. It would be useful to further test the algorithms under real high latitude data.
3. The SAKF2 predefined models could be further enriched in order to include additional scenarios (i.e. for the case of receivers subject to high dynamics or including more scintillation levels).
4. It should be underlined that KF algorithms require a high computation burden due to the several multiplications necessary to compute the KF gains at each iteration. A way to reduce this computational cost could consist in storing the KF gains in a look up table and selecting the suitable gains for the specific scenario. This approach could be particularly advantageous for systems with limited computational power often employed by commercial GNSS receivers.

5. The focus of this research work was to improve the robustness of the carrier tracking stage, and specifically of the PLL, which is particularly sensitive to high dynamics and deep fading produced by scintillation. The assessment of the proposed algorithms has been performed by evaluating the tracking performance and the scintillation monitoring capabilities of the propose architectures. Since more robust PLL algorithms will affect the quality of the final position solution, the impact of the proposed tracking schemes on the positioning performance should be investigated.

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Appendix A: The Fresnel Zone

The purpose of this appendix is to clarify the concept of Fresnel zone. The strength of ionospheric scintillation is determined by the size of the irregularity, the distance between receiver and irregularity, and the first Fresnel zone. As illustrated in Figure A-1, the Fresnel zone can be defined by ellipsoids with the transmitter and the receiver as foci.

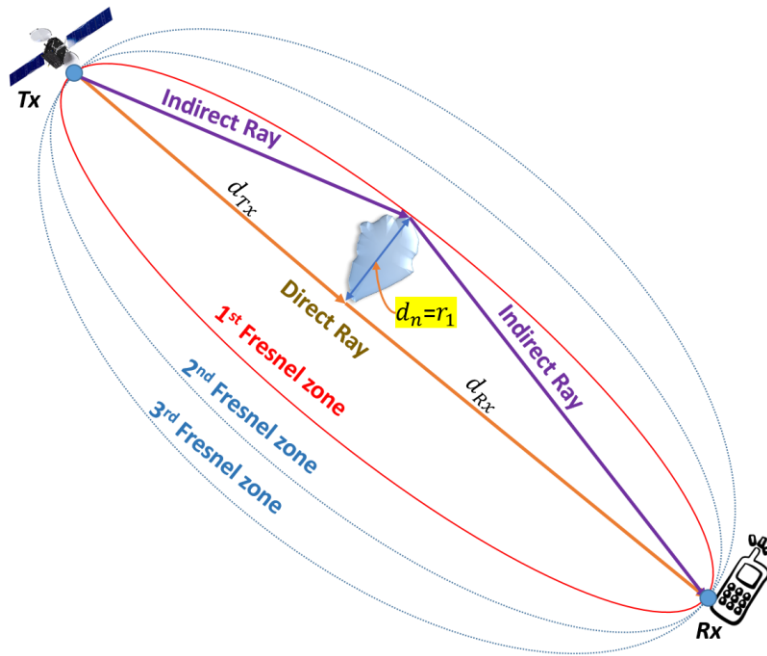


Figure A- 1: Graphical representation of the Fresnel zone.

The surface points of these ellipsoids are locations where the difference between the length of the direct and the indirect path is a multiple of a half wavelength of the transmitted signal, as expressed by the following expression

$$d_{IR} = d_{DR} + n\lambda/2 \quad (A.1)$$

Where d_{IR} is the distance of the indirect ray, d_{DR} is the distance of the direct ray, the integer n defines the different ellipsoids and λ is the signal wavelength. When the (A.1) is valid the transversal dimension of the irregularity, indicated in Figure as d_n represents a Fresnel zone radius. If the distance between the transmitter and the irregularity, defined as d_{Tx} is so that $d_{Tx} \gg d_n$, d_n can be approximated by the following

$$d_n \cong \sqrt{\frac{n\lambda d_{Tx} d_{Rx}}{d_{Tx} + d_{Rx}}} \quad (A.2)$$

where d_{Rx} is the distance between the irregularity and the receiver. When the integer n is equal to one d_n indicates the first Fresnel zone radius d_1 . The irregularities with transversal dimension equal to the first Fresnel zone radius are the ones which more affect the signal propagation. Indeed the amplitude scintillation for irregularities with dimensions bigger than the first Fresnel zone radius can be neglected (Knight, 2000).

Appendix B: Software receiver overview

The algorithms proposed by this thesis have been implemented by modifying an open source software receiver developed by Borre et al. (2007). The main advantage of a software receiver is its flexibility enabling rapid modifications and testing of different algorithms. Figure B-1 represents the structure of the above software receiver which starts after the front end and includes the acquisition, tracking and PVT stages. This software receiver processes the IF data provided by an external front-end, which is the only hardware unit, as described in Chapter 6. In Figure B-1, the modifications performed to the software receiver are represented in pink while the original algorithms are in grey.

Indeed, as part of the research different modifications have been implemented to update the original tracking algorithm which is based on a second order PLL with fixed loop bandwidth and time of integration. Specifically, a third order PLL which is more suitable to cope with the dynamics produced by ionospheric scintillation, has been included into the software receiver and other advanced tracking architectures have been implemented. They are the novel tracking algorithms proposed by this thesis, namely SAKF1 and SAKF2, a standard KF based tracking scheme and two different versions of Adaptive KFs tracking algorithms. Details on the above tracking architectures can be found in Chapters 5 and 6. Moreover, the software receiver has been integrated with a module in charge of monitoring the signal quality and the tracking performance through the estimation of different parameters, such as the C/N_0 , the Phase Lock Indicator (PLI), the phase and

Doppler jitter. The latter parameters are defined in Chapter 6. Finally a scintillation monitoring block has been implemented. This module provides a number of scintillation indices (S_4 , Φ_{60} , p , T) to quantify the level of ionospheric scintillation. Details on the computation of these indices can be found in Chapters 4 and 6.

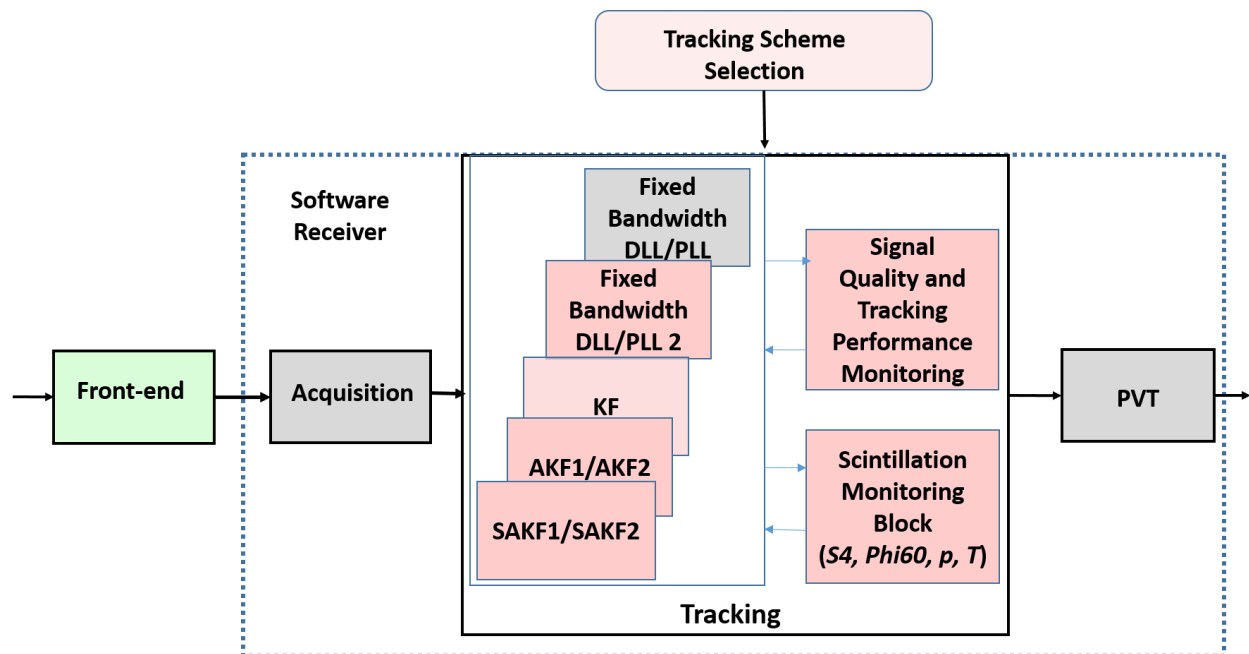


Figure B- 1: Software receiver overview. The original modules of the software receiver are represented in gray while the new implemented modules are represented in pink.