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Contents

Introduction	iv
1 Evolution of the power systems	1
1.1 Wide Area Monitoring Protection and Control systems	4
1.2 Key components of WAMPAC systems	4
1.2.1 Protective and measuring transformer	5
1.2.2 Phasor Measurement Unit	8
1.2.3 Phasor Data Concentrator	9
1.2.4 Control center and actuators	10
1.2.5 Communication network	11
1.3 Challenges in WAMPAC system design and implementation . . .	13
2 Algorithms for synchrophasor estimation in presence of high dynamics and transients	16
2.1 Models of electrical signals and synchrophasor measurements . .	16
2.2 Estimation errors and synchrophasor standard	20
2.3 Phasor Measurement Unit algorithms	27
2.4 Taylor-Fourier models for PMU algorithms	31
2.5 Left and right Taylor-Fourier estimates	35
2.5.1 Criteria for selecting the Taylor-Fourier expansion	37
2.5.2 Blending left and right estimates	38
2.6 Tests and results	41
2.6.1 Proposed methods configuration	41
2.6.2 Steady-state and dynamic tests	45
2.6.3 Amplitude and phase step tests	50
2.6.4 Analysis of computational burden	59
2.6.5 Response time comparison of state-of-art PMU algorithms	61
3 Advanced synchronization technologies	63
3.1 Overview of synchronization in power systems	63
3.1.1 Global Navigation Satellite System	64

3.1.2	Synchronization time protocols	69
3.1.3	White Rabbit protocol	72
3.1.4	Examples of time-sensitive power system applications . .	75
3.2	Implementation of WR architectures	76
3.3	Characterization of synchronization errors	79
3.4	Examples of the contribution to measurement error of synchro- nization in power system applications	83
3.4.1	Harmonic synchrophasor measurement	83
3.4.2	Traveling wave-based fault location	85
4	Time quality assessment in all-digital measurement chain	88
4.1	Digital substation architecture	88
4.2	Fully digital measurement chain	91
4.3	Proposal of Sampled Value PMU with time quality evaluation functionality	94
4.4	Implementation of the measurement setup	96
4.5	Test and results of the implemented SV-PMU	99
4.5.1	Time synchronization sensitivity of SV-PMU	99
4.5.2	Results of SV-PMU time quality evaluation	102
5	Latency measurement architectures for WAMPAC systems	106
5.1	Latency challenges in WAMPAC systems	107
5.2	Contributions of the overall WAMPAC latency	107
5.3	Proposed latency characterization system	111
5.3.1	WAMPAC Characterizer	112
5.3.2	WAMPAC Emulator	116
5.4	Tests and results of WAMPAC latency measurements	121
5.4.1	Test scenarios and setup	121
5.4.2	Metrological validation of the WAMPAC Characterizer . .	125
5.4.3	Experimental tests on real WAMPAC systems	131
	Conclusions	137
	Appendix	141
	Bibliography	144
	List of Figures	163
	List of Tables	167
	Acronyms	168

Acknowledgements	172
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Introduction

The power grid is undergoing a significant transformation as it evolves from a traditional centralized system to more decentralized, dynamic, and smart networks. This transition is driven by the increasing integration of distributed generation, Renewable Energy Sources (RESs), Electric Vehicles (EVs), and other electrified loads, all of which present new challenges in terms of stability, reliability, and control of the power system. In this context, Wide Area Monitoring, Protection, and Control (WAMPAC) systems have emerged as an interesting solution to enhance the real-time observability and responsiveness of modern power grids.

WAMPAC systems take advantage of synchronized measurement architectures to provide a comprehensive view of grid dynamics over large geographical areas. By utilizing data from Phasor Measurement Units (PMUs), WAMPAC systems enable improved detection of grid disturbances, faster response times to system events, and optimized grid management. These advancements are crucial for ensuring the security and efficiency of future power systems, especially in light of the increasing complexity and variability introduced by distributed generation based on RESs.

This thesis focuses on the development and validation of key measurement tools and methodologies for WAMPAC systems, addressing the core challenges of latency, synchronization accuracy, and synchrophasor estimation even in presence of dynamic and transient conditions.

The thesis is structured as follows:

- Chapter 1 explores the evolution of power systems and introduces the role of WAMPAC in modern power grids. It also outlines the key components of WAMPAC systems and the challenges associated with their design and implementation.
- Chapter 2 delves into algorithms for synchrophasor estimation, proposing and evaluating new methods based on Taylor-Fourier models to minimize the measurement error even in presence of real dynamic and transient conditions.

- Chapter 3 investigates advanced synchronization technologies that can improve WAMPAC-based monitoring, such as the White Rabbit protocol, and their impact on two key power system applications.
- Chapter 4 addresses the assessment of time quality in PMUs, focusing on the functionality of fully digital measurement chains and proposing a method to evaluate the quality of the synchronization at run time to extend the all-digital PMU operation.
- Chapter 5 presents architectures for measuring the latency of WAMPAC systems, with a detailed discussion of the proposed latency characterization system, including both laboratory validation and real measurements from the field.
- Conclusions provide a summary of the results achieved and a discussion of potential future developments in WAMPAC systems for electric power grids.

By providing novel insights into architectures, instruments and measurement algorithms, this thesis aims to contribute to the ongoing development of reliable and efficient WAMPAC systems for modern power networks.

Chapter 1

Evolution of the power systems

The global power systems landscape has experienced a significant transformation over the past decades. Traditional power grids, designed around centralized power generation from large synchronous generators, are gradually being replaced by more complex and dynamic systems. As mentioned in the introduction, this evolution is primarily driven by the decarbonization need and the growing world electrical energy demand, by the integration of Renewable Energy Sources (RESs), the advancements in technology, and the increasing demand for more reliable and resilient power grids in this more variegated context.

Indeed, over the past century, global primary energy consumption, which is still mainly based on the use of fossil fuels, has grown exponentially, going beyond five times the past use in order to be able to sustain economic and population growth [1]. Energy production from fossil fuels is among the main contributors to greenhouse gas emissions, widely recognized as causing significant environmental and climate impacts [2]. It is clear that the energy model on which the planet's growth of recent years has been built is no longer sustainable, thus imposing a global commitment to a progressive and as rapid as possible decarbonization and efficiency in all energy sectors, especially in the electric one.

One of the key pillars of global decarbonization is the concept of “Electrification of Everything”. This refers to the transition towards powering almost all sectors of the economy, such as transportation, industry, and residential heating, through electricity rather than fossil fuels. Electrification is seen as a crucial strategy for reducing carbon emissions, especially when electricity is generated from RESs, such as solar and wind energy [3].

A prominent example of electrification is the widespread adoption of Electric Vehicles (EVs), which, while reducing the carbon footprint of transportation, presents new challenges for the electric grid. Indeed, the increasing number of EVs requires the expansion of charging infrastructure and poses the risk of introducing substantial peaks in electricity demand, particularly during times of simultaneous charging, which could strain grid capacity [4].

These transformations are not only of technical nature but also reflect a broader shift toward global sustainability goals, such as the 17 Sustainable Development Goals (SDGs) outlined in the United Nations 2030 Agenda for Sustainable Development [5]. In line with the SDGs, the modernization of power systems plays a vital role in supporting Goal 7 (Affordable and Clean Energy), which emphasizes the importance of increasing access to clean energy and enhancing the share of renewables in the global energy mix. Additionally, these advancements contribute to Goal 11 (Sustainable Cities and Communities) by enabling resilient urban energy systems that reduce emissions and improve air quality. Supporting Goal 12 (Responsible Consumption and Production), the transition towards cleaner energy sources aligns with the goal of sustainable resource management, helping to reduce fossil fuel dependency. Finally, Goal 13 (Climate Action) underscores the urgency of mitigating climate change impacts, which is fundamentally supported by transitioning to a cleaner and more resilient power grid based mainly on RESs that reduces greenhouse gas emissions.

At the same time, the integration of RESs into the power grid introduces additional complexities. Renewable energy generation is inherently variable, depending on weather conditions, and often does not align with peak electricity demand. This mismatch creates a need for advanced grid management strategies and large-scale energy storage systems to ensure a reliable and continuous power supply [6]. Moreover, RESs are typically operated in direct current (DC) mode and require dedicated power converters to be interconnected to the rest of the grid [7]. Such interface relies on fast-switching power electronics that do not contribute to the overall system inertia [8], [9] and inject spurious contributions in the harmonic and supraharmonic range [10], [11]. It is thus reasonable to expect that the power signals are going to be subject to higher distortion rates and faster dynamics. In general, this scenario makes the system more prone to sudden contingencies or even blackouts (as proven by the recent events in California [12] or South Australia [13]).

All these power system evolutions, are schematized in Fig. 1.1, which clearly highlights the increase in complexity from the traditional centralized power grids to the modern decentralized ones.

This transition also presents several challenges, including:

- Voltage and frequency regulation: The variability of RESs and the increasing number of Distributed Energy Resources (DERs) make it more difficult to maintain stable voltage and frequency levels, increasing the risk of instability and grid blackout.
- Energy storage and peak load management: Large-scale energy storage solutions are critical for balancing supply and demand, particularly during periods when renewable generation is insufficient to satisfy peak demand.
- Grid congestion: As more electrified devices such as EVs are connected to the grid, local distribution networks could become congested, leading to potential bottlenecks in power flow.
- Cybersecurity risks: The increasing reliance on digital infrastructure for grid management makes the power system more vulnerable to cyberattacks, which could disrupt operations and compromise reliability.

The evolving electric grid will have to overcome all these challenges to achieve a fully sustainable system without reducing its efficiency.

As the power system changes towards greater electrification and decarbonization, it must become also more flexible, resilient, and capable of real-time monitoring and control. This transformation relies heavily on the implementation of the so-called Wide Area Monitoring, Protection, and Control (WAMPAC) systems.

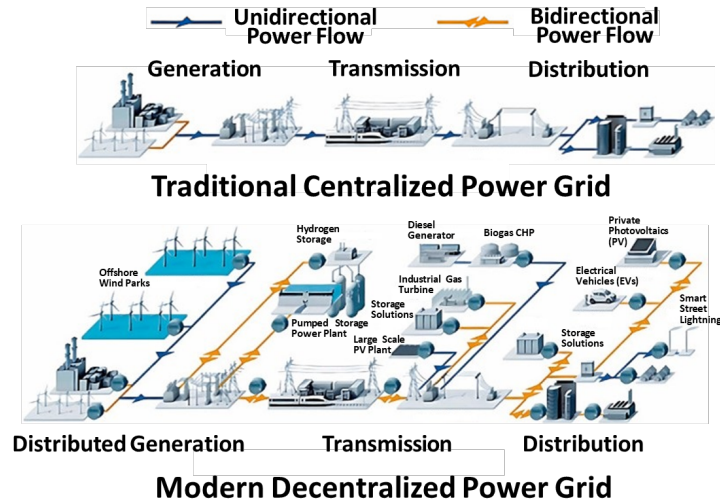


Figure 1.1: Typical schematic of traditional centralized and modern decentralized electric power grids [14].

1.1 Wide Area Monitoring Protection and Control systems

In this context, WAMPAC systems have emerged as an innovative solution to enhance the real-time observability and responsiveness of modern power grids. Indeed, they provide the real-time data and control capabilities necessary to protect the wide area system, maintaining grid stability, optimizing the performance and ensuring that the grid can support the increasing demand for clean, renewable energy.

The main functions of WAMPAC systems include monitoring, protection, and control of a large area of the network. Monitoring is typically accomplished through synchronized measurement instruments, which allow operators to detect disturbances and anomalies on the network with adequate speed and accuracy to solve the possible problem before it propagates in the system. The protection aspect of WAMPACs is to ensure network integrity by quickly isolating failed network sections, minimizing the risk of cascading failures. Control functions, on the other hand, enable rapid automated corrective actions such as load shedding, voltage and frequency regulation, and dynamic reconfiguration of the network to prevent instability, maintaining stability and reliability.

WAMPAC systems are currently being designed and implemented mainly for transmission networks, where they improve the coordination of protection and control mechanisms over large geographical areas. However, systems with similar features and based on similar solutions are also expected to be implemented in distribution networks, given the increasing complexity and dynamic behavior of these systems [15]. For these reasons, WAMPAC systems will be explored in detail, both for their current application scenario in transmission systems and their future potential in distribution networks.

1.2 Key components of WAMPAC systems

The WAMPAC is the evolution of the traditional Wide Area Monitoring Systems (WAMSs), well established in the last few decades, and for this reason key monitoring components are common. Indeed, these systems involve, as the first element in their monitoring architecture, the Phasor Measurement Unit (PMU) [16], which represents the “sensor” of the grid for the capability to measure the electrical quantities of interest [17], and the Phasor Data Concentrator (PDC), which enables the reception and time alignment of measurements from PMUs distributed throughout the power grid and sends the data to another PDC or to a control center [18]. The PMU, for its proper operation, is typically preceded

by the protective or measuring transformers, that can attenuate the level of voltage and current to be measured.

These key tools are presented in a typical diagram of a WAMPAC system in Fig. 1.2. A key feature specific to WAMPAC systems, as highlighted in this diagram, is the Actuator. Unlike WAMSs that focus solely on monitoring and reporting grid states, WAMPAC systems incorporate actuators to take corrective actions on the electric power grid based on the data received, such as adjusting generation. These actuators are responsible for automating responses to grid disturbances, making WAMPAC capable of real-time protection and control.

All these key tools will be described and discussed one by one in the following sections.

1.2.1 Protective and measuring transformer

Protective and measuring transformers, generically indicated as Instrument Transformers (ITs), play a crucial role in every power system, as they convert high-voltage and high-current signals to lower levels that can be safely acquired by protective relays and metering devices [20]. These transformers include Current Transformers (CT) and Voltage Transformers (VT), sometimes also called potential transformers, according to the signal to be converted. The quality and accuracy of the conditioned signals provided by these transformers have a direct impact on the measurements used by protection and monitoring functionalities of WAMPAC systems [21].

In particular, protective transformers are designed to operate reliably during fault conditions, ensuring that protective relays receive conditioned signals even in extreme situations such as short circuits or overcurrent events. The protection aims to isolate faulted sections of the grid, preventing further damage and ensuring the stability of the power system [22]. Protective transformers are classified into accuracy classes such as 5P and 10P, where “P” stands for protection, and the number indicates the maximum percentage error at the rated current or voltage. These classes ensure that protection systems receive sufficiently reliable signals during high-current and high-voltage events to trigger appropriate responses during abnormal conditions.

In contrast, measuring transformers focus on delivering highly accurate measurements under normal operating conditions, which are essential for tasks like system monitoring and grid management. For this reason, accuracy is crucial in these transformers, and they are categorized into accuracy classes such as Class 0.1, Class 0.2, and Class 0.5, which represent the maximum percentage errors of the transduced voltage magnitude or current allowed during steady-state conditions. This level of accuracy, which is significantly better than protective trans-

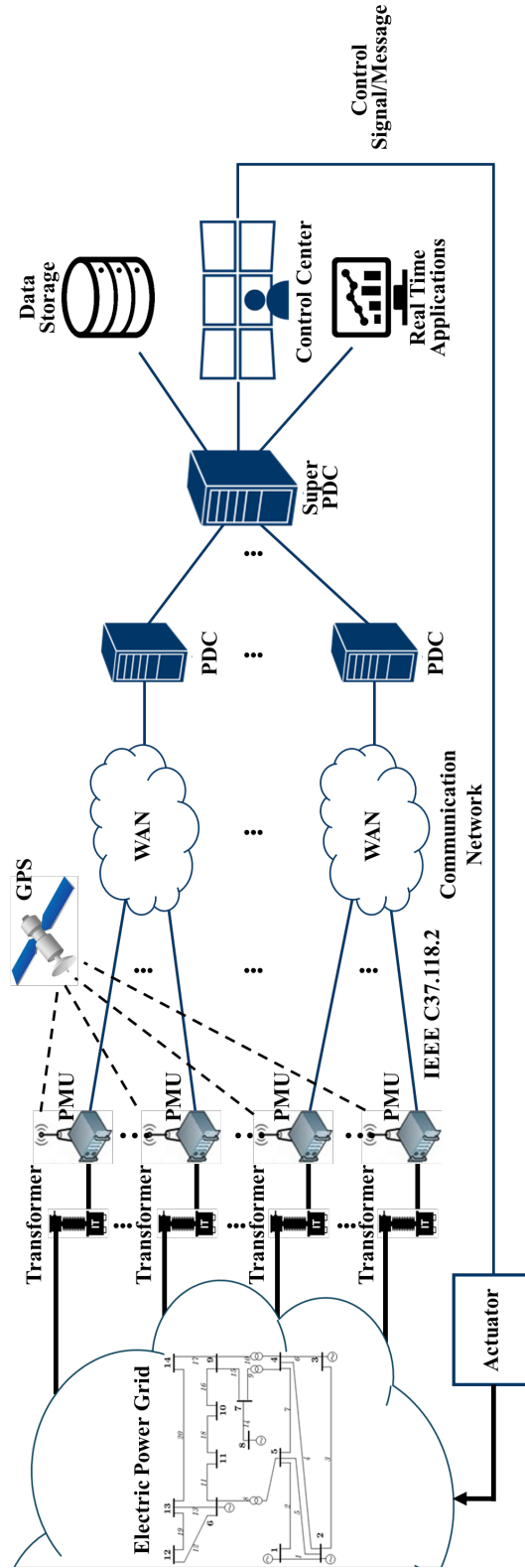


Figure 1.2: Diagram of a generic WAMPAC system, composed of the electric power grid (for instance IEEE 14-bus system [19]), instrument transformers, PMUs, PDCs, control center and actuator.

formers, is necessary for advanced grid operations, where the decision-making process is based on accurate measurements [23] and thus gives a quality limit to WAMPAC systems input data.

Fig. 1.3 shows a typical connection between the CT and VT, connected to the high-voltage lines, and the PMU, where the burden is also present.

An important innovation in the field of ITs is the development of Low Power Instrument Transformers (LPITs). Compared with conventional inductive transformers, these electronic transducers generally provide lower voltage levels in output and have low power consumption. If the LPITs still provide analog outputs at the secondary side, these signals can be converted into digital form by Stand Alone Merging Units (SAMUs), which allow seamless integration into digital substations and eliminate the need for analog-to-digital (A/D) conversion in PMUs [25].

This process is further supported by IEC Std 61850 [26], which constitutes the framework for the development of fully digital and interoperable substations, where different Intelligent Electronic Devices (IEDs) communicate using common protocols and data formats [27]–[29]. In particular, IEC 61850-9-2 introduces the Sampled Value (SV) protocol for the transmission of the secondary output of LPITs in digital form [30]. SAMUs convert the analog output of LPITs into digital SV streams, which are then published on the substation network. In this way, the SV are available not only for storage and post-processing applications in the control room, but also for an online estimation process as performed by IEDs [28], such as PMUs [31], [32] or Power Quality Instruments (PQIs) [33],

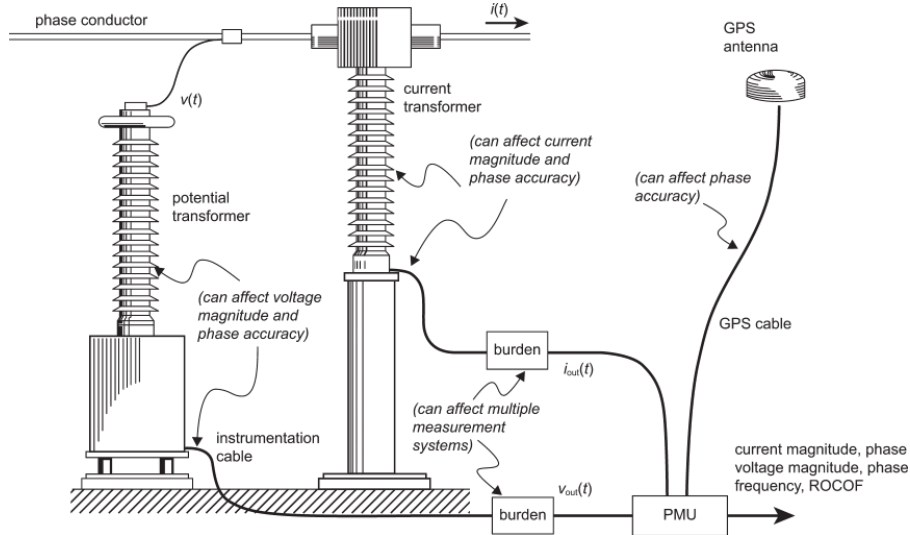


Figure 1.3: Typical connection diagram between transformers and PMU [24].

[34].

The fully digital measurement chains and the role of SAMUs within it will be discussed in more detail in Chapter 4. Instead, the next section will describe in detail the PMU and its importance within a WAMPAC system.

1.2.2 Phasor Measurement Unit

The PMU, as mentioned above, is the key element of WAMPAC systems that allows measuring phasors for voltages and currents with an accurate time reference, namely the so-called synchronized phasors or synchrophasors, instantaneous frequency and rate of change of frequency (ROCOF) [17]. Therefore, measurements made by PMUs require a common and absolute time reference in order to be compared with each other, and, for this reason, synchronization technologies that can provide high temporal accuracy (better than $1\text{ }\mu\text{s}$ [17]) are used. To this aim, PMUs are often equipped with a Global Positioning System (GPS) receiver or make use of packet-based synchronization, like that provided by Precision Time Protocol (PTP) [35], to tag each measurement with an absolute timestamp in the Coordinated Universal Time (UTC) timescale. The measurements computed by PMUs are usually encapsulated in data frame packets compliant with the IEEE C37.118.2 Standard [36]. PMUs perform measurements at a specific reporting rate (RR), typically in the order of tens of frames per second (fps), and the one most commonly used by transmission system operators (TSOs) is 50 fps, which means a packet containing measurements every 20 ms [37].

PMUs are distributed in the power network (e.g., in 380 kV and 220 kV substations) and can be configured according to two compliance classes, Protection Class and Measurement Class (in brief P- and M-class respectively), defined by the synchrophasor standard IEC/IEEE Std 60255-118-1 [17] (PMU Std in what follows). The PMU Std also defines the corresponding requirements under both steady-state and dynamic conditions in terms of estimation accuracy, response time and reporting latency for the two classes (these terms will be clarified in Section 2.2). These requirements will be explained in detail in Chapter 2 together with the description of some PMU algorithms and the characterization of the two proposed P and M algorithms. In general, P-class is used for applications that require “fast” measurements, i.e. with short response times, and thus it is expected to play a role especially for electric system protection, whereas M-class has less stringent timings and is preferable where better disturbance rejection and larger off-nominal ranges are more important. For this reason, the use of P-type algorithms is often considered preferable for WAMPAC systems, in order to reduce the so-called reporting latency introduced by PMUs, which

mainly depends on the algorithm computation window. An interesting solution for WAMPAC systems would be to choose PMUs that meet the stringent requirements of the two classes in a unique PMU algorithm, often referred to as “P + M” [38].

For accurate state estimation, the optimal placement of PMUs in the grid is also important, balancing between estimation accuracy and overall cost of the measurement infrastructure [39], [40].

PMUs play a crucial role in a wide range of WAMPAC applications, as the monitoring and control actions taken automatically using some policy in the control center are based on synchrophasor, frequency and ROCOF measurements. In the literature, there are several examples of WAMPAC policies that use these measurements. For instance, in [41], WAMPAC applications based on ROCOF estimation with three established state-of-the-art algorithms are analyzed, and their impact on the successful grid restoration is shown. In [42], several WAMPAC applications are described, including recording of dynamic events, real-time system state determination and phase-angle monitoring (PAM). PAM, for example, is an application of significant relevance to WAMPAC systems, as it provides information about power transfer [43]. In addition, it allows handling angular instability issues in the transmission network by activating real-time protection functions in order to prevent propagation phenomena and thus improve resilience [44]. As often happens with WAMPAC applications, there are still no standardized recommendations, e.g., regarding angular difference bounds for transmission grids, as pointed out in [44]. Indeed, the thresholds considered by WAMPAC depend on the specific context and TSO expertise.

The ability to measure accurate synchrophasors with high temporal accuracy and compare these measurements across the grid is critical to both existing and future applications. As WAMPAC systems evolve, many other applications are expected to emerge, all of which will rely on the output provided by PMUs or PMU-like devices.

1.2.3 Phasor Data Concentrator

The other key component in the measurement chain of the WAMPAC system is the PDC. As mentioned above, PDCs are in charge of receiving synchronized data from scattered PMUs or other PDCs and aggregating those with identical timestamps. This time-aligned data can be sent as an outgoing data stream to devices with a variety of functions including storage, display, monitoring, protection and control. In addition, since packets from different PMUs and/or PDCs may not arrive at the same time, PDCs wait a maximum time called “wait

time” for the arrival of data from all connected devices, after which the packet with partial data is sent anyway [18]. This prevents a high increase in latency. In this context, PDC operations must be performed in real-time, mitigating the PDC latency as much as possible to meet the real-time requirements of WAMPAC applications. For example, in [45], the architecture and validation of a low-latency PDC are studied using different network infrastructures, e.g., with fiber-optic cables or with 4G LTE wireless technology.

The PDC for its basic function can also be lacking of external time synchronization, performing time alignment using only the timestamp of the packet itself. However, it is becoming increasingly common to use synchronization in this device as well to increase its functionality. Indeed, this element, initially conceived as a module responsible for receiving data from the PMUs, is now assuming an increasingly active role in control systems, as it is the first element to supervise a network portion in the synchrophasor hierarchy. Indeed, the IEEE C37.247-2019 PDC Standard [18] recommends the ability of PDCs to implement control logic, such as event detection and classification, and examples of these techniques are described in [46]. Other examples of PDCs with advanced capabilities can be found in [47], where a prototype PDC capable of handling delay and accurately evaluating PMU data flow latencies is presented.

In many WAMPAC architectures, the PDC is directly connected to a control center where aggregated data from the entire wide area network is analyzed to verify grid stability and intervene in the system if necessary.

1.2.4 Control center and actuators

The control center in WAMPAC architectures uses the synchronized data provided by PMUs and PDCs to perform real-time analysis and decision-making. Moreover, it is responsible for activating protection schemes, adjusting generation or load, and managing power flows to maintain grid stability [48]. In cases of system instability or fault detection, the control system can automatically trigger corrective actions such as isolating affected grid sections or activating backup generation resources [49]. Additionally, load shedding mechanisms can be automatically triggered by the control center in case of a major system imbalance, helping prevent widespread outages [50].

Advanced control algorithms, often integrated with machine learning and predictive analytics, are being developed to improve the responsiveness and accuracy of WAMPAC systems [51]. These algorithms can process vast amounts of real-time data, predict potential grid disturbances, and optimize the grid performance under varying operational conditions [52]. The combination of advanced control systems with real-time data from PMUs ensures that WAMPAC systems

provide a robust solution to the challenges posed by modern power grids.

The actuators within WAMPAC systems play a critical role in executing the decisions made by the control center (sent with specific control signal or message in the network that will be explained in detail in Chapter 5), directly interacting with the physical grid to implement corrective measures. These actuators interface with various grid components for instance to control voltage levels, adjust power generation, and stabilize the grid. For example, tap changers in transformers are common actuators used to regulate the voltage in power transmission lines. By adjusting the transformer's tap settings, the control system can maintain the voltage within specified limits, ensuring optimal power delivery [53]. Moreover, the control actuators are used in Flexible Alternating Current (AC) Transmission Systems, the so-called FACTSs, such as Static Var Compensators (SVCs) and Static Synchronous Compensators (STATCOMs), which are used to dynamically manage reactive power flow, stabilizing the grid and improving power quality [54].

The control system and the actuators are not specifically detailed here nor discussed from here on for two main reasons: the thesis focuses on measurement aspects and, unlike the monitoring system, which relies on standard or well-defined elements such as PMUs and PDCs, the design and functionality of the control system largely depend on the choices made by the operator. One example of a real control system will be explored in detail in Chapter 5.

1.2.5 Communication network

The communication infrastructure connecting PMUs, PDCs, and control center is another key component of WAMPAC systems. These networks must provide high-speed, reliable data transfer to ensure that WAMPAC systems can respond to grid events in real-time. WAMPAC systems can rely on dedicated communication networks to avoid congestion and delays caused by other non-critical traffic [55]. These networks must be designed with redundancy and fault tolerance to ensure continuous operation, even during network failures or cyberattacks [56].

Cybersecurity is a growing concern for WAMPAC systems, as their reliance on digital infrastructure makes them even more vulnerable to potential attacks [57]. In this context, cybersecurity in WAMPACs involves several levels, from design to system vulnerability testing. A detailed discussion of the topic can be found in [58], and a series of technical recommendations are reported in [59]. Focusing on specific examples, in [60] possible attacks on a WAMPAC infrastructure equipped with attack-resilient algorithms and a defense architecture are explored. A concrete example are time synchronization attacks that can change

the UTC time provided by synchronization, impacting on the accuracy of phase angle and derived measurements and also on the associated timestamps, thus affecting even the sending time and latency of PMU data [61]–[63].

The communication network between PMUs and PDCs can introduce delays in system response time, which is why minimizing communication delays is a priority for WAMPAC systems. The choice of communication technology affects the overall latency, reliability, and scalability of the WAMPAC system [64]. Several communication technologies are available for these systems, each with its own advantages and limitations:

- Ethernet is an extremely widely used communication technology worldwide due to its high-speed data transfer capabilities and wide compatibility. Ethernet networks offer typical bandwidths ranging from 100 Mbps to 10 Gbps, with a typical range of 100 meters for twisted pair cabling (which can be extended using switches or fiber) [65]. The reliability and speed of Ethernet are particularly advantageous for ensuring fast real-time data transmission. However, Ethernet networks often require extensive cabling, which can increase installation costs and complexity, especially in large-scale network deployments.
- Fiber-optic communication is highly favored in WAMPAC systems because of its ability to transmit data over long distances with minimal latency and interference. Fiber optic networks support extremely high bandwidths, from 1 Gbps to 100 Gbps, and can cover distances up to 100 km without signal degradation, making them ideal for handling the large volumes of data generated by PMUs in real time, especially in wide-area network applications [66]. In addition, this technology is often used to transmit the White Rabbit (WR) synchronization protocol, which, as will be discussed in detail in Chapter 3, requires this technology to ensure high temporal accuracy. In addition to the advantage of high bandwidth, optical fiber is immune to electromagnetic interference, ensuring data integrity even in high-noise environments such as substations. However, the installation costs of fiber optic networks are relatively high, especially in remote or difficult-to-access areas.
- Wireless communication technologies such as Wi-Fi, cellular networks (4G/5G) and ZigBee offer the flexibility of installation without the need for physical cabling. Wireless networks typically have bandwidth capacities ranging from 10 Mbps to 1 Gbps, with communication ranges depending on the technology. For example, Wi-Fi typically covers up to 100 meters, while 4G/5G networks can provide communications for several kilometers

[67]. Wireless systems are useful in areas where wired infrastructure is impractical or too expensive to install. However, wireless communication is more susceptible to signal degradation, interference, and security vulnerabilities than wired alternatives, which can lead to higher latency and lower reliability in some conditions.

- Power line communication (PLC) uses existing power lines to transmit data, providing a cost-effective communication solution that does not require additional wiring. PLC technology can provide bandwidths ranging from 100 kbps to 200 Mbps, depending on implementation, with communication ranges from hundreds of meters to several kilometers [67]. Although this can be advantageous in locations where laying new communication lines is difficult or expensive, PLC is subject to noise and interference from the power lines themselves, which can degrade signal quality and increase latency. In addition, the limited bandwidth of the PLC makes it less suitable for transmitting large amounts of data over large areas.

Each of these communication technologies has its own strengths and trade-offs, and the selection of the appropriate technology for WAMPAC systems depends on the specific needs of the network, including the distance between PMUs and PDCs, the volume of data being transmitted, and the required response times.

To reduce the delay caused by communication, the network topology must also be carefully evaluated [68], [69], and different communication technologies, like supplementary wireless link, can be used to compensate temporary wired network failures [70]. For instance, in [71], the experience of the Croatian Transmission System Operator with a PMU-based angle stability protection application is presented, discussing the communication latency due to the different geographical locations and communication technologies of the PMUs installed in their network. The various contributions to WAMPAC latency will be explained in detail in Chapter 5.

1.3 Challenges in WAMPAC system design and implementation

The implementation of WAMPAC systems presents several technical and operational challenges that need to be addressed to ensure their effectiveness in modern power grids. The thesis focuses on the measurement-related challenges and proposes solutions to some of the most relevant, as illustrated below.

One of the primary challenges is the design and implementation of PMU

algorithms capable of handling fast dynamics and abrupt transients, which are increasingly characteristic of modern power systems. Traditional synchrophasor estimation algorithms may struggle to accurately capture the rapid changes in voltage and current that occur during such events. To address this gap, more advanced algorithms are being designed to provide more accurate synchrophasor, frequency, and ROCOF measurements during rapid changes preserving performance also in steady-state condition of the electric signals. This issue will be discussed in detail in Chapter 2, where new algorithms based on Taylor-Fourier models are proposed and evaluated to minimize errors in dynamic and transient conditions.

Time synchronization is also a major concern. Indeed, PMUs require accurate time synchronization across the grid to provide accurate synchrophasor measurements, as even small errors (larger than 1 μ s) in time synchronization can lead to significant phase, frequency and ROCOF errors. Although Global Navigation Satellite System (GNSS) technology is commonly used for this purpose, its reliability can be compromised by satellite visibility, environmental factors or attacks, potentially resulting in loss of synchronization. In addition, new advanced synchronization technologies, such as WR, are emerging as excellent substitutes for this traditional synchronization method. Despite these advancements, it remains essential to minimize synchronization errors and their contribution to the measurement errors to avoid performance degradation in power system applications. For this reason, Chapter 3 investigates these synchronization technologies, focusing especially on emerging WR technology, and their impact on power system applications.

A further challenge is the assessment of time quality in WAMPAC context, which is particularly important as power grids move towards fully digital measurement chains. In these systems, the accuracy of time synchronization has a direct impact not only on PMUs but also on the devices responsible for digitizing the analog signal, i.e. the SAMUs. To address this problem or, at least, to tackle potential degradations, there is a need for innovative methods to assess the quality of measurement synchronization. These methods are essential to ensure that timing information in digital power systems is sufficiently reliable for real-time WAMPAC applications and, for this purpose, Chapter 4 will discuss the proposal of a novel method for evaluating synchronization quality at the PMU side in these contexts, and thus extend all-digital PMU operating conditions.

Finally, measuring latency in WAMPAC systems remains another key challenge. WAMPAC relies on real-time data to detect network events and respond quickly, but the delay of each device in the network can introduce latency that affects the system's ability to react in real-time. Accurate measurement and

characterization of the overall WAMPAC latency and its constituent contributions, such as the latency of PMUs, PDCs, communication networks, control chain, etc., are critical, particularly in highly interconnected networks, to ensure that latency does not compromise overall system performance. This challenge will be thoroughly addressed in Chapter 5, where a latency characterization system is proposed, including laboratory validation and real measurements from the field.

As mentioned above, all these challenges will be addressed in detail in dedicated chapters, proposing innovative solutions for each of them that might contribute to the improvement of WAMPAC systems in the near future and help WAMPAC design evolution.

Chapter 2

Algorithms for synchrophasor estimation in presence of high dynamics and transients

In this chapter, the model of electrical signals in dynamic conditions and the synchrophasor estimation problem statement will first be discussed in general. After that, the various components that constitute a Phasor Measurement Unit (PMU) will be described, focusing especially on the synchrophasor estimation algorithm. Afterwards, the PMU algorithms based on the Taylor-Fourier model will be presented, before introducing two proposed algorithms compliant with the P- and M-class of the PMU Std [17], which allow to effectively cope with abrupt transients.

2.1 Models of electrical signals and synchrophasor measurements

In alternating current (AC) power systems during ideal operating conditions, the voltage and current signals can be modeled mathematically in time domain as a waveform characterized by a single sinusoidal tone with parameters constant in time as:

$$x(t) = A_0 \cdot \cos(2\pi f_0 t + \varphi_0) \quad (2.1)$$

where t is the time in seconds, A_0 the cosine peak-amplitude, φ_0 the initial phase angle of the signal and f_0 the nominal frequency of the power system (i.e. 50 or 60 Hz depending on the country), corresponding to the angular frequency $\omega_0 = 2\pi f_0$. This model is strictly valid just during ideal conditions of the AC

power system, but in the reality, a power system is never in nominal steady-state condition. For this reason, the parameters of Eq. (2.1) are rarely time-invariant and typically exhibit various dynamic behaviors, especially in modern power systems. Furthermore, purely sinusoidal conditions are an abstraction. Frequency fluctuations are certainly the most obvious phenomena and are typically related to changes in load-generation imbalances, to the inertia of large generators and to the automatic speed controls of the generators [48]. Analogously, the amplitude and phase angle of the fundamental component of voltage or current are also affected by dynamics or transient phenomena that can be relatively slow, as in the case of amplitude and phase oscillations, typically characterized by frequencies up to 3 Hz in Low-Frequency Oscillation (LFO) [72], [73] or 14 Hz as in forced oscillations [74], or faster as in the case of switching events or faults. The latter conditions usually produce changes in voltage and current waveforms with spectral components of even several hundreds of kilohertz [17], which can thus be modeled as steps. These events can also produce even larger frequency fluctuations than forced oscillations [75]. Furthermore, other components of different nature and at different frequencies can overlap to the fundamental component, including:

- Harmonics, i.e. spectral components at frequencies that are integer multiples of the instantaneous AC system frequency which, as mentioned above, can be different from the nominal frequency f_0 . These phenomena are typically produced by non-linearities in the system and their frequency is usually less than a few kilohertz. The literature on PMUs usually considers the effects of harmonics up to the 50th, assuming that higher frequency components are attenuated, for instance by the PMU's anti-aliasing filter, and are also too far from the fundamental component (the measurand) in the frequency spectrum to impact significantly on the measurement.
- Interharmonics are spectral components at frequencies that are not integer multiples of the system frequency. Sometimes the terms supraharmonic and subharmonic are also used, which simply indicates a special case of interharmonic for frequency components, respectively, much higher or lower than the power system frequency. These phenomena are usually caused by non-linear interactions in power system, such as those introduced by the fast switching actions of the Renewable Energy Source (RES) power converters [76], not synchronized with the fundamental frequency, or by non-linear loads as the arc furnaces that draw current at irregular intervals or at frequencies different from that of the fundamental [77].
- Wideband noise, which includes both “grid noise”, such as thermal noise, and “measurement noise” which is added by any of the measurement equip-

ment in the acquisition (e.g. digitization error). Both sources are usually modeled as white uniform or Gaussian noise processes with zero mean.

For the above reasons, Eq. (2.1) can be rewritten by creating a generalized and more complete signal model as follows:

$$x(t) = A_1(t) \cdot \cos(\psi_1(t)) + d(t) \quad (2.2)$$

being $A_1(t)$ the instantaneous peak-amplitude, $\psi_1(t)$ the instantaneous phase angle and $d(t)$ the disturbance signal containing additive contributions. The terms $A_1(t)$ and $\psi_1(t)$ can be expressed with different models to account for different dynamics in the signal. For instance, considering $f_1(t)$ the instantaneous frequency variable over time, which is typically close to the nominal frequency of the power system f_0 , and φ_1 the initial phase angle of the fundamental component, $\psi_1(t)$ can be modeled as:

$$\psi_1(t) = 2\pi \int_0^t f_1(\tau) d\tau + \varphi_1 \quad (2.3)$$

where the integral $\int_0^t f_1(\tau) d\tau$ takes into account the cumulated phase up to time t . All the considered quantities are continuous functions of time, as indicated by the t -dependent notation.

Taking into account the interference signals mentioned above, the $d(t)$ term in Eq. (2.2) can be modeled as follows:

$$\begin{aligned} d(t) = & \sum_{h \in \mathcal{H} \subset \mathbb{N}, h > 1} A_h(t) \cos\left(2\pi \int_0^t f_h(\tau) d\tau + \varphi_h\right) + \\ & + \sum_{ih=1}^{N_{ih}} A_{ih}(t) \cos\left(2\pi \int_0^t f_{ih}(\tau) d\tau + \varphi_{ih}\right) + \epsilon(t) \end{aligned} \quad (2.4)$$

In Eq. (2.4), the first two terms model the contributions of any superimposed sine tone, thus harmonics and interharmonics excluding the main tone. Each harmonic (interharmonic) is characterized by a peak amplitude $A_h(t)$ ($A_{ih}(t)$), an instantaneous frequency $f_h(t)$ ($f_{ih}(t)$), and an initial phase angle φ_h (φ_{ih}) with therefore instantaneous phase $\psi_h(t) = 2\pi \int_0^t f_h(\tau) d\tau + \varphi_h$ ($\psi_{ih}(t) = 2\pi \int_0^t f_{ih}(\tau) d\tau + \varphi_{ih}$). For harmonics $f_h(t) = hf_1(t)$. Under normal operating conditions, the peak amplitude of harmonics can range up to percentage points with respect to the amplitude A_1 , depending on the presence of nonlinear loads such as rectifiers, inverters, and switching devices, and usually decreases with increasing harmonic order [76]. Interharmonics, in practical scenarios, are characterized by amplitudes generally lower than those of harmonics, but they can occasionally reach levels comparable to low-order harmonics, particularly in systems with significant nonlinear loads or variable-speed drives [77], [78]. Instead, for initial phase, it is difficult to make assumptions, about the harmonics and interharmonics initial phase, which may not necessarily match

that of the fundamental φ_1 . In the PMU Std [17] these components are usually tested one at a time, but, in reality, there are often multiple tones together in the electric signal. Finally, the last element $\epsilon(t)$ represents broadband noise and can also include any other contributions not considered with the previous terms.

To simplify the analysis of electrical systems in the sinusoidal steady-state, the phasor transformation has been historically adopted. This transformation can be employed if and only if the instantaneous frequency and the peak amplitude are time-invariant, i.e., the system is in steady-state. By using it, sinusoidal functions can be mapped into complex numbers. In particular, this transformation makes it possible to represent a sinusoidal time function such as the one in Eq. (2.1) with a single complex constant. To derive it, Eq. (2.1) can be rewritten as:

$$x(t) = \sqrt{2} \Re \{ \bar{X}_0 e^{j2\pi f_0 t} \} \quad (2.5)$$

where $\bar{X}_0$ is the phasor (overline denotes complex-valued quantities) defined as:

$$\bar{X}_0 = \frac{A_0}{\sqrt{2}} e^{j\varphi_0} = X_0 (\cos \varphi_0 + j \sin \varphi_0) = X_{0,r} + jX_{0,i} \quad (2.6)$$

in which the magnitude $X_0 = \frac{A_0}{\sqrt{2}}$ is the root mean square (RMS) value, while $X_{0,r}$ and $X_{0,i}$ are the real and imaginary parts of the complex value, called rectangular coordinates. The phasor can also be represented with polar coordinates as $X_0 \angle \varphi_0$, defined by the RMS amplitude and the initial phase angle and, as in Eq. (2.6), the transformation from one type of coordinate to the other can be done using the Euler identity $e^{j\varphi_0} = (\cos \varphi_0 + j \sin \varphi_0)$. In Eq. (2.5), $\Re\{\}$ is the real operator that permits to obtain the real part of the phasor. In what follows, the imaginary operator $\Im\{\}$ will also be used, which, analogously, provides the imaginary part of a complex number.

As mentioned above, the assumption of stationarity is not valid in the real power system, so the concept of phasor cannot be applied directly to the measurement. However, even if the system is not stationary, the changes in voltage and current can be assumed to be relatively slow. With this quasi-stationary approximation, phasor analysis can still be applied, introducing the dynamic phasor and using it for complex signal models as Eq. (2.2). The dynamic phasor, according to PMU Std [17], is defined by extending the traditional phasor concept to include time-varying components as:

$$\bar{X}_1(t) = X_1(t) e^{j\varphi_1(t)} = X_1(t) e^{j\psi_1(t) - 2\pi f_0 t} \quad (2.7)$$

where $X_1(t) = \frac{A_1(t)}{\sqrt{2}}$ is the RMS value of the instantaneous magnitude and $\varphi_1(t)$ is the phasor phase angle defined as the phase difference between the instantaneous angular position $\psi_1(t)$ and the phase due to the nominal frequency f_0 .

To compare phasors at different positions in the network, due to the dependence of the instantaneous phase on time, there is the necessity of a common time reference, as mentioned in Chapter 1. For this reason, in the PMU context, the measured phasors are referred to Coordinated Universal Time (UTC) time and are called synchrophasors.

As mentioned above, synchrophasor estimation involves calculating the magnitude $X_1(t)$ and phase angle $\varphi_1(t)$ of the fundamental component of the AC waveform in the power grid at specific reporting instants.

In addition to synchrophasor, instantaneous frequency and rate of change of frequency, called ROCOF by the PMU Std [17], can be defined with the following formulas:

$$f_1(t) = \frac{1}{2\pi} \frac{d\psi_1(t)}{dt} = f_0 + \frac{1}{2\pi} \frac{d\varphi_1(t)}{dt} \quad (2.8)$$

$$\text{ROCOF}(t) = \frac{df_1(t)}{dt} = \frac{1}{2\pi} \frac{d^2\psi_1(t)}{dt^2} = \frac{1}{2\pi} \frac{d^2\varphi_1(t)}{dt^2} \quad (2.9)$$

In Eq. (2.8), it is clear how the frequency can be measured by computing the deviation with respect to the nominal frequency (second term of the formula), while Eq. (2.9) indicates ROCOF as the first derivative of the instantaneous frequency.

2.2 Estimation errors and synchrophasor standard

To evaluate the performance of PMU algorithms, various indices have been defined to quantify deviations in the synchrophasor, frequency, and ROCOF estimates and they will be described in what follows.

The Amplitude Error (AE) quantifies the deviation between the actual RMS amplitude X_1 of the phasor and the estimated RMS amplitude $\widehat{X}_1$. From now on, the symbol $\widehat{\cdot}$ will be used to denote estimates. This error can be expressed in relative value as in the following formula (in percentage points):

$$\text{AE}(n) = \frac{|X_1(n) - \widehat{X}_1(n)|}{X_1(n)} \times 100\% \quad (2.10)$$

where the n refers to the index of the measurement (the continuous time t is replaced by the discrete sequence of reporting instants¹).

The Phase Error (PE) measures the difference between the actual phase angle $\varphi_1(n)$ and the estimated phase angle $\widehat{\varphi}_1(n)$. It is defined (and usually in

¹Here the reporting interval is not reported for the sake of simplicity.

centiradians) as:

$$\text{PE}(n) = \varphi_1(n) - \hat{\varphi}_1(n) \quad (2.11)$$

The Total Vector Error (TVE) is a comprehensive index that takes into account both the amplitude and phase errors. It is defined as the relative Euclidean distance between the estimated and actual synchrophasors as (in percentage points):

$$\text{TVE}(n) = \frac{\sqrt{\left(\hat{X}_{1,r} - X_{1,r}(n)\right)^2 + \left(\hat{X}_{1,i}(n) - X_{1,i}(n)\right)^2}}{\sqrt{X_{1,r}^2(n) + X_{1,i}^2(n)}} \times 100\% \quad (2.12)$$

TVE is widely used as a metric to assess the overall performance of synchrophasor estimation algorithms, as it incorporates both magnitude and phase deviations, but obviously, by including both, the contribution of AE and PE within it is lost. Therefore, in Section 2.6 AE and PE errors will also be presented alongside TVE.

The Frequency Error (FE) assesses the accuracy of the frequency estimate $\hat{f}_1$ compared to the actual system frequency f_1 as:

$$\text{FE}(n) = f_1(n) - \hat{f}_1(n) \quad (2.13)$$

and it is usually expressed in mHz.

The ROCOF Error (RFE) is used to quantify the accuracy of the estimate $\widehat{\text{ROCOF}}$ against the actual value of ROCOF. It can be computed in Hz/s as:

$$\text{RFE}(n) = \text{ROCOF}(n) - \widehat{\text{ROCOF}}(n) \quad (2.14)$$

In the PMU Std, compliance tests for PMUs are divided into steady state and dynamic state. Each type of test to be satisfied requires accuracy requirements in terms of TVE, absolute FE and absolute RFE that are different based on the P- or M-class.

Steady-state tests are defined as conditions in which the instantaneous parameters of the reference signals, including also the parameters of the interfering signals, are constant throughout the duration of each test and are varied only from one test to another. These tests, including those dedicated to signal frequency and magnitude variation, harmonic distortion and out-of-band (OOB) interference, will be described one by one in what follows.

During the signal frequency tests, the reference signal is composed of a single sinusoidal component with a frequency f_1 that varies by 0.1 Hz between each test and constant amplitude. The frequency range to be tested depends on the compliance class and in P-class the frequency deviates at most by ± 2 Hz, while in M-class by $\pm f_{RR}/5$ Hz if $10 \text{ Hz} \leq f_{RR} < 25 \text{ Hz}$ otherwise by ± 2 Hz if $f_{RR} < 10 \text{ Hz}$ or ± 5 Hz if $f_{RR} \geq 25 \text{ Hz}$, where f_{RR} is the reporting rate

(RR) expressed in Hz². Therefore, in M-class, the maximum frequency range is between 45 and 55 Hz, which corresponds to the typical case of RR at 50 frame per seconds (fps).

In the signal amplitude tests (separate for voltage and current), similar to the previous test, the only parameter to vary is the amplitude A_1 . This value, for P-class, can vary between 80 % and 120 % of rated voltage and between 10 % and 200 % of rated current, when considering voltage and current signals, respectively. In case of M-class, A_1 varies between 10 % and 120 % for the voltage and between 10 % and 200 % for the current³.

The harmonic distortion tests evaluate the impact of harmonics on the quality of the synchrophasor, frequency and ROCOF estimated by the PMU. In this tests, the fundamental signal in steady-state is corrupted by a single harmonic at a time, characterized by an amplitude equal to 1 % or 10 % of the fundamental amplitude A_1 in case of P- or M-class compliance, respectively. The test is repeated by varying the harmonic index h from 2 to 50.

Finally, the OOB interference test allows to evaluate the influence of interharmonics on the quality of the estimates and is defined only for M-class compliance. OOB compliance shall be tested with a main tone signal characterized by a frequency between $f_0 - 0.1f_{RR}/2$ and $f_0 + 0.1f_{RR}/2$ (for example between 47.5 and 52.5 Hz if $f_1 = f_0 = 50$ Hz and $RR = 50$ fps) and the other parameters are constant. Single interharmonics shall be superimposed on this main tone with an amplitude up to 10 % of A_1 and a frequency between 0 and $f_0 - f_{RR}/2$ or between $f_0 + f_{RR}/2$ and $2f_0$, therefore up to the 2nd harmonic (if $f_{RR} < 100$ and $f_{RR} \geq 10$).

In dynamic tests, the instantaneous parameters of the main tone of the reference signal, i.e., its amplitude $A_1(t)$ and frequency $f_1(t)$, are not constant over the duration of each test. The PMU Std does not refer to the dynamic behavior of any interfering signals, which are not included in dynamic test conditions. The dynamic tests defined by the PMU Std are amplitude and phase modulation, frequency ramp, and amplitude and phase-angle step tests.

In all these tests, the terms $A_1(t)$ and $\varphi_1(t)$ of Eq. (2.2) are expressed with different models to account for the specific test condition.

The amplitude modulation (AM) and phase modulation (PM) tests aim to evaluate the quality of PMU estimates during independently modulated variation of the instantaneous amplitude $A_1(t)$ and phase $\psi_1(t)$ of the main tone of the reference signal and are defined as follows:

$$A_1(t) = A_0 [1 + k_{am} \cos(2\pi f_m t)] \quad (2.15)$$

² $1/RR$ is thus the reporting interval in seconds.

³Rated values at PMU input are typically 70 V RMS or 120 V RMS for voltage, and 5 A RMS for current.

$$\psi_1(t) = 2\pi f_0 t + k_{pm} \cos(2\pi f_m t - \pi) \quad (2.16)$$

In Eq. (2.15), k_{am} is the amplitude modulation factor, i.e., a scalar that defines the depth of the AM ($k_{am} = 0$ means that no modulation occurs and the signal retains its constant amplitude A_0), and f_m represents the modulation frequency, that chooses how fast the amplitude will be oscillating over time. Analogously, in Eq. (2.16), the k_{pm} term is the phase angle modulation factor, which determines the magnitude of phase change and f_m denotes the frequency of the PM. In this case, the resulting frequency of the signal will exhibit oscillations from the nominal frequency f_0 due to phase oscillation.

In both AM and PM tests, f_m must be varied with a step of at least 0.2 Hz between each test, within a range from 0.1 to the maximum modulation frequency f_M , i.e., $\min(f_{RR}/10; 2)$ and $\min(f_{RR}/10; 5)$ in the case of P-class or M-class compliance, respectively. Thus, for the typical case of $f_{RR} = 50$ Hz, f_M reaches 2 Hz for P-class and 5 Hz for M-class. Instead, the modulation factors k_{am} and k_{pm} can be kept constant at 0.1 (maximum variations of $\pm 10\%$ in amplitude from the nominal value A_0) and 0.1 rad, respectively.

The frequency ramp test allows testing the linearity of the PMU filter and the correction of any PMU delays in the timestamp. In this test, the signal is modeled with a linear frequency ramp by:

$$\psi_1(t) = 2\pi f_0 t + \pi R_f t^2 \quad (2.17)$$

where R_f is the frequency ramp rate in Hz/s (i.e. the ROCOF value during ramp) that corresponds to a constant value and $t = 0$ is the ramp start. In Eq. (2.17), since $\psi_1(t)$ is a second-order function of time, the instantaneous frequency, in fact, becomes a linear function of time. It should be reminded that the frequency depends on the first derivative of the phase as already seen in Eq. (2.8).

For this test, the PMU Std prescribed a frequency ramp rate $R_f = \pm 1$ Hz/s (positive or negative ramps) and all other parameters of the fundamental are maintained constant. The frequencies to be tested have a range between $f_0 - 2$ Hz and $f_0 + 2$ Hz for P-class compliance and between $f_0 - \min(f_{RR}/5; 5)$ Hz and $f_0 + \min(f_{RR}/5; 5)$ Hz in the case of M-class compliance. As in the steady-state case, the frequency can vary with a maximum range of $45 \div 55$ Hz with a typical R_f of 1 Hz/s. Before and after the frequency ramp, the test requires the signal to be in steady-state. For this reason, to not include the transition from the starting steady-state frequency to the ramping frequency and analogously from the ramping frequency to the ending steady-state frequency, the PMU Std prescribed a time interval called “exclusion interval” $t_{\text{exclusion}}$, where the measurements made by PMU under test are excluded from the evaluation. This interval is defined as $2/f_{RR}$ for P-class and $7/f_{RR}$ for M-class compliance, which

CHAPTER 2. ALGORITHMS FOR SYNCHROPHASOR ESTIMATION IN PRESENCE OF HIGH DYNAMICS AND TRANSIENTS

corresponds, respectively, at 40 ms and 140 ms for $f_{RR} = 50$ Hz.

Finally, the amplitude and phase step tests are aimed at testing the response of PMUs to sudden changes. The PMU Std defined $A_1(t)$ and $\psi_1(t)$ as:

$$A_1(t) = A_0 (1 + k_{as}u(t)) \quad (2.18)$$

$$\psi_1(t) = 2\pi f_0 t + k_{ps}u(t) \quad (2.19)$$

where $u(t)$ is the unit step function, k_{as} and k_{ps} is the magnitude and phase step size, respectively. The PMU Std proposed to describe $u(t)$ with an instantaneous step, defining it as:

$$u(t) = \begin{cases} 0, & t < t_{\text{step}} \\ 1, & t \geq t_{\text{step}} \end{cases} \quad (2.20)$$

being, t_{step} the time instant at which the step occurs. Increasingly complicated models can adjust $u(t)$ by including, for example, the possibility of instantaneous amplitude and phase steps or with a rise characterized by more realistic transient, e.g., linear as in the case presented in Section 2.6.3 or transients described by functions with exponential growth as the Heaviside function used in [79].

The PMU Std requires testing steps with both positive and negative variations of amplitude ($\pm 10\%$) and phase angle ($\pm 10^\circ$), using Eq. (2.18) and (2.19). In addition to the indexes described before, the PMU Std introduced other indexes, such as response time, delay time and overshoot, in order to evaluate better the transition between two steady-state measurements, before and after a step change. Indeed, for these tests, PMU Std does not provide stringent accuracy levels but allows for exceeding the steady-state requirements during a time interval called “response time”. The response time T_R is determined as the difference between the time instant when the measurement goes out of a specified accuracy limit and the time when it re-enters and remains within that limit in case a positive or negative step (of amplitude or phase angle) is applied to the input of the PMU. For this reason, this index allows to understand for how long the PMU measurements have to be considered “invalid” during a step and it can be evaluated not only for TVE, but also for FE and RFE errors (considering the corresponding limits under steady-state conditions).

Response time differs for TVE, FE and RFE, which are henceforth referred to respectively as $T_{R,\text{TVE}}$, $T_{R,\text{FE}}$ and $T_{R,\text{RFE}}$, and the prescribed limits are summarized in Table 2.1. For example, in the case of f_0 and f_{RR} equal to 50 Hz, the response time limits for TVE, FE, and RFE become 40, 90, and 120 ms for P-class and 140, 280, and 280 ms for M-class, respectively.

The delay time measures the time interval between the instant in which the step is applied and the instant in which the measured quantity (it depends on

CHAPTER 2. ALGORITHMS FOR SYNCHROPHASOR ESTIMATION IN PRESENCE OF HIGH DYNAMICS AND TRANSIENTS

Table 2.1: Response time requirements for amplitude and phase step tests of P- and M-class compliance

Test	$T_{R,TVE}$ [ms]		$T_{R,FE}$ [ms]		$T_{R,RFE}$ [ms]	
	P	M	P	M	P	M
Amp./ph. step	$2/f_0$	$7/f_{RR}$	$4.5/f_0$	$\max(14/f_{RR}, 14/f_0)$	$6/f_0$	$\max(14/f_{RR}, 14/f_0)$

the type of step) reaches 50 % of the final value after the transition. This index is useful to verify that any delays introduced by the PMU processing have been adequately compensated. The requirements imposed for delay time are the same for the two compliance classes and are equal to $1/(4f_{RR})$.

The overshoot/undershoot is the maximum/minimum value of the peak of the measurement curve with respect to the true value applied, considering the quantity that undergoes a step variation. Finally, the limit for maximum overshoot and undershoot is equal to 5 % for P-type algorithms and 10 % for M-type algorithms.

In Fig. 2.1, an example of amplitude measurement and TVE trend is shown for a 10 % amplitude step. A graphical representation of TVE response time, of the delay time and of the overshoot is also reported.

Table 2.2 summarizes the steady-state and dynamic compliance requirements of PMU Std in terms of TVE, FE and RFE for both classes at $f_{RR} = 50$ Hz. Accuracy limits for amplitude and phase step tests are not mentioned in the table since it refers to the limits of steady-state tests (e.g., signal frequency) for response time computation. From this table, it can be seen that M-class requires more stringent accuracy limits except in the cases of AM and PM, where the requirements are quite larger since the modulation frequency has significantly wider ranges than in P-class. Accuracy requirements may also depend on RR as, for example, in the case of AM and PM, the limits of $|FE|$ can be computed in mHz as $30 \cdot \min(f_{RR}/10; 2)$ and $60 \cdot \min(f_{RR}/10; 2)$ for P-class and M-class, respectively. The formulas for deriving the requirements of the various tests of the PMU Std for each class with different f_{RR} and f_0 values will not be discussed in detail here, as they will not be used in the following. Please refer to PMU Std for more detail [17].

Another important parameter for evaluating the PMU is the reporting latency, which is the time delay between the moment when the packet containing the PMU measurement data is available at the output and the reporting instant timestamp t_r associated with the measurements of the packet itself. As mentioned, this index will be discussed in detail in Chapter 5 as a contribution to the total latency of a WAMPAC system.

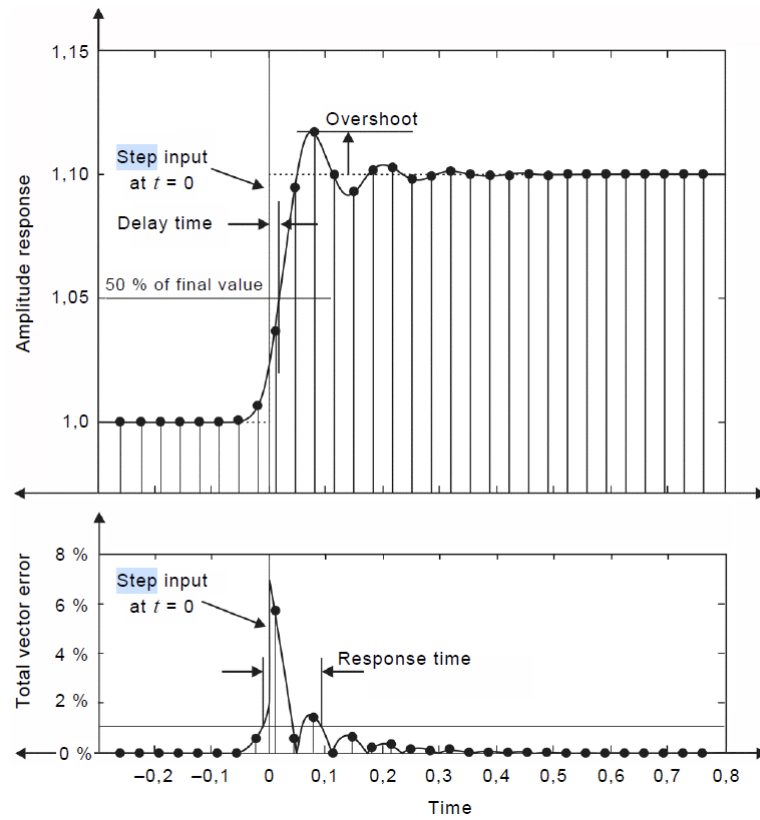


Figure 2.1: Example of the PMU Std for the amplitude measurement and the TVE of a 10 % amplitude step at $t = 0$.

CHAPTER 2. ALGORITHMS FOR SYNCHROPHASOR ESTIMATION IN
PRESENCE OF HIGH DYNAMICS AND TRANSIENTS

Table 2.2: Accuracy requirements for steady-state and dynamic compliance tests of P- and M-class with $f_{RR} = 50$ Hz

Test	TVE [%]		FE [mHz]		RFE [Hz/s]	
	P	M	P	M	P	M
Signal frequency	1	1	5	5	0.4	0.1
Signal magnitude	1	1	5	5	0.4	0.1
Harmonic distortion	1	1	5	5	0.4	0.1
OOB interference	-	1.3	-	10	-	-
Amplitude modulation	3	3	60	300	2.3	14
Phase modulation	3	3	60	300	2.3	14
Frequency ramp	1	1	10	10	0.4	0.2

2.3 Phasor Measurement Unit algorithms

Building a PMU is a complex task that involves competencies in signal processing, digital architecture design, time synchronization, etc. Indeed, a PMU device is composed of several elements, as shown in Fig. 2.2:

- V and I analog channels: receive analog voltage and current signals from the grid. These signals are previously attenuated by ITs which, as already mentioned, scale the voltage and current levels of the grid to manageable levels for the PMU.
- Anti-aliasing filter: before the analog signals are converted into digital form, they may pass through anti-aliasing filters, which are essential to prevent aliasing. Aliasing occurs when higher-frequency components of the signal, beyond the Nyquist frequency (half of the sampling rate), are misinterpreted as lower-frequency components during the sampling process. This can alter the samples and result in incorrect phasor, frequency, and ROCOF estimates. The anti-aliasing filter removes these frequency components to ensure that only the relevant signal within the desired bandwidth is passed through for digitization. Proper design and implementation of these filters are crucial for maintaining the integrity of synchrophasor measurements, especially for dynamic conditions [80].
- Sample and Hold (S&H): the filtered analog signals are then captured and temporarily holded sample by sample in the S&H circuit at regular time intervals. This step is essential to discretize the analog signal in time, maintaining the amplitude constant during the brief sampling period (note that with a typical sampling frequency $f_s = 10$ kHz the sampling period corresponds to $T_s = 100$ μ s) needed for analog-to-digital (A/D) conversion, ensuring that the A/D converter can accurately process the sampled value without distortion or variation caused by fluctuations in the signal.

- A/D converter (ADC): the analog sampled signals are then quantized by the ADC circuit. This process discretizes the sampled signal in amplitude by mapping the voltage and current waveforms to a finite set of digital levels at regular intervals. For PMUs to accurately capture the synchronized phasors of voltage and current across multiple phases, it is essential that each channel is sampled simultaneously to avoid timing discrepancies between channels, which could otherwise introduce phase errors and affect the accuracy of synchrophasor measurements. To achieve this goal, PMUs often use synchronized S&H circuits between the voltage and current channels and a single ADC, increasing the hold phase, or multiple ADCs to convert the signals in parallel but at the same time, increasing the cost of the instrument. Acquisition through S&H and ADC is often regulated by internal PMU clocks, that are aligned with the time synchronization module using a Phase Locked Loop (PLL) control circuit. The accuracy of the A/D conversion is critical to preserve the fidelity of the original signal and ensure accurate measurements, and for this reason Successive Approximation Register (SAR) converters or Sigma-Delta converters capable of providing high number of bits (up to 24 bits or slightly more) are often used. However, Sigma-Delta converters may introduce an inherent processing delay that can be challenging to compensate for. Therefore, the choice of converter must be carefully considered by choosing on the basis of sampling rate, number of bits, and accuracy the one best suited for the specific case of the application.
- Time synchronization: This block provides a UTC time reference to be associated with the measurements and, depending on the PMU design, to lock the sampling process. The implementation depends on the synchronization source used, which as mentioned earlier can be Global Positioning System (GPS), a Precision Time Protocol (PTP) server, or, more advanced, a White Rabbit (WR) source. For example, for classical GPS synchronization, PMUs often have an integrated GPS receiver, which decodes the space signal from available satellites, computes the time offset and thus provides an accurate UTC reference. These concepts on synchronization will be discussed in more detail in Chapter 3.
- Synchrophasor estimation algorithm: this unit is the heart of the PMU that performs the synchrophasor estimation and thus calculates synchrophasor, frequency, and ROCOF measurements with the associated UTC timestamp. The algorithms are often implemented in microprocessors or Field Programmable Gate Arrays (FPGAs) to ensure low computation latency.
- Data communication system: Once the measurement data has been pro-

cessed, this unit builds the packet compliant with the IEEE-C37.118.2 Standard [36] and sends the packet to the network through the adopted data communication technology.

Most of the literature regarding PMU focuses on the design and analysis of synchrophasor, frequency, and ROCOF estimation algorithms, as the quality of PMU estimation comes mainly from a well-formulated algorithm that takes into account as much as possible the different sources of error considering the various operating conditions (and tests of the PMU Std). In general, PMU estimation algorithms can be divided into two categories: static and dynamic algorithms [81].

The former assume that the waveform parameters are constant within the adopted observation window (set of samples to process), and thus use a static signal model. Obviously, these algorithms often have high errors in the presence of dynamic signals due to their nature and are often insufficient to achieve compliance with the standard classes. Static algorithms are often based on direct implementation of the Discrete Fourier Transform (DFT). Interpolated DFT (IpDFT) is commonly used to improve the accuracy of DFT measurements by estimating the frequency of the fundamental through interpolation techniques, which is particularly useful in the case of noncoherent sampling (estimation of signals with a non-integer number of cycles within the observation window) [82], [83]. Examples of advanced IpDFT algorithms can be found in [84]–[88]. The fundamentals of several of them will be explained briefly in Section 2.6.5. Other algorithms in literature also use zero-crossing methods [89], [90], adaptive filters [91], [92], compressive sensing algorithms [93], [94], and many others.

In contrast, dynamic algorithms also include the possibility of time-varying waveform parameters within the measurement window, and thus use a dynamic signal model. Some of the dynamic estimation algorithms are based on the Taylor-Fourier approach [95]–[101], whose basic principle will be described in detail in Section 2.4.

In the literature, a still open issue of both categories is represented by PMU response to step changes in magnitude and phase angle [102]. Such steps model fast transients and can be interpreted as sudden transitions between two steady-state conditions. Such discontinuities can be hardly represented with slowly-modulated sinusoidal signal models, which are those traditionally employed in many PMUs [103]. A possible solution consists in applying a step detection technique and adapting the estimation algorithm accordingly [97]. This approach was also adopted in [104] to switch between a P-class and an M-class measurement channel and in [105] to define an algorithm that is simultaneously compliant with classes P and M.

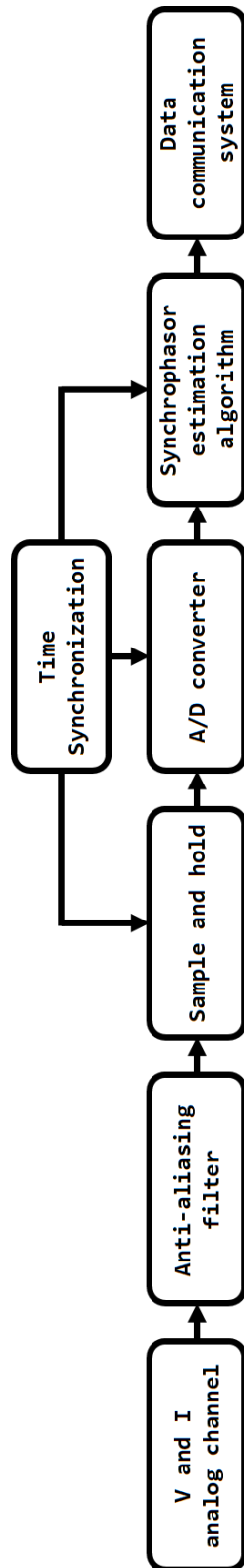


Figure 2.2: Typical block diagram of a standalone PMU.

By means of the Hilbert Transform (HT), it is possible to define the underlying analytic signal and check either its envelope, like in [106] and [107], where step location is intended for calibration purposes, or its spectrum [86]. An alternative solution relies on the use of dictionaries of pre-defined sets of waveforms, either inspired by the PMU Std test conditions [108] or by suitably modified wavelet functions [109]. However, both these solutions suffer from different shortcomings. The first one requires a precise and effective method for the computation of the analytic signal associated with the considered window [110]. The second one is computationally demanding (i.e., not easily implementable in real-time applications) and depends on the dictionary granularity [111].

For these reasons, in this chapter the focus is on a new PMU estimation algorithm that, depending on the configuration, has been proven to be compliant with P- and M-class requirements. More precisely, a Taylor-Fourier Multifrequency (TFM) model [98] that includes the fundamental component and its first four harmonic terms is considered as the starting point to propose a new algorithm that improves its performance in the presence of steps. TFM represents an extension of the Taylor-Fourier filters introduced in [100], where it was proposed to track the fundamental component oscillations. In the literature, the response of TFM to amplitude and phase steps has also been extensively analyzed [112]. By properly setting the expansion degrees, a bank of Taylor-Fourier filters can be also employed to promptly detect and classify fault events [113].

2.4 Taylor-Fourier models for PMU algorithms

To explain the foundations of TFM models, it is better to first introduce the Taylor-Fourier (TF) basic approach. Considering the signal model in Eq. (2.2) and an observation window having length $T_w = MT_s$ with $M = 2N + 1$ samples and centered on t_r , a generic sample $x(t_r + nT_s)$ is expressed through the Taylor expansion of the synchrophasor (and its complex conjugate) around t_r , truncated to the K_1 -th degree as:

$$\begin{aligned} x(t_r + nT_s) = \sum_{k=0}^{K_1} \frac{(nT_s)^k}{k!} \left[\frac{\bar{X}_1^{(k)}(t_r)}{\sqrt{2}} e^{j\omega_0(t_r+nT_s)} + \right. \\ \left. + \frac{\underline{X}_1^{(k)}(t_r)}{\sqrt{2}} e^{-j\omega_0(t_r+nT_s)} \right] + r(t_r) \end{aligned} \quad (2.21)$$

where $\bar{X}_1^{(k)}$ is the k -th order derivative of the synchrophasor (and $\bar{X}_1^{(0)} = \bar{X}_1$), while underline indicates complex conjugation and $r(t_r)$ is the residual vector. Using matrix notation, the M -size vector of the samples (collected with sampling

interval T_s) is written as:

$$\mathbf{x}(t_r) = \begin{bmatrix} x(t_r + NT_s) \\ \vdots \\ x(t_r) \\ \vdots \\ x(t_r - NT_s) \end{bmatrix} = \bar{\mathbf{B}}\bar{\mathbf{p}}(t_r) + \mathbf{r}(t_r) \quad (2.22)$$

having introduced the M -size residual vector $\mathbf{r}(t_r)$ and the $(2K_1 + 2)$ -size vector of the model parameters

$$\bar{\mathbf{p}}(t_r) = \left[\bar{X}_1^{(0)}(t_r) \cdots \bar{X}_1^{(K_1)}(t_r) \quad \underline{X}_1^{(0)}(t_r) \cdots \underline{X}_1^{(K_1)}(t_r) \right]^\top \quad (2.23)$$

All of these parameters refer to the instant t_r , even if from here on it will be often omitted for the sake of brevity. The $M \times (2K_1 + 2)$ matrix $\bar{\mathbf{B}}$ can be partitioned as

$$\bar{\mathbf{B}} = \begin{bmatrix} \bar{\Phi}(\omega_0)\mathbf{A}(K_1) & \Phi(\omega_0)\mathbf{A}(K_1) \end{bmatrix} \quad (2.24)$$

where

$$\bar{\Phi}(\omega) = \text{diag} \left[e^{j\omega NT_s} \dots e^{-j\omega NT_s} \right] e^{j\omega t_r} \quad (2.25)$$

is the $M \times M$ phase rotation matrix, while

$$\mathbf{A}(K) = \frac{\sqrt{2}}{2} \begin{bmatrix} 1 & NT_s & \frac{(NT_s)^2}{2} & \dots & \frac{(NT_s)^K}{K!} \\ \vdots & \vdots & \vdots & & \vdots \\ 1 & 0 & 0 & & 0 \\ \vdots & \vdots & \vdots & & \vdots \\ 1 & -NT_s & \frac{(-NT_s)^2}{2} & \dots & \frac{(-NT_s)^K}{K!} \end{bmatrix}. \quad (2.26)$$

is a $M \times (K + 1)$ matrix.

An estimate $\hat{\mathbf{p}}(t_r)$ of the TF model parameters can be obtained by minimizing the Euclidean norm of the misfit, namely the deviation between the samples in the current time window and their reconstructions obtained from the model. If the number of parameters defining the TF model (namely $2(K_1 + 1)$) is lower than M , the optimization problem can be solved in the least squares (LS) sense by computing $\bar{\mathbf{H}} = \bar{\mathbf{B}}^\dagger$, with † denoting the Moore-Penrose inverse. The m -th element of $\hat{\mathbf{p}}$ is obtained by convolving the samples with the m -th row of $\bar{\mathbf{H}}$, thus its elements correspond to the (complex-valued) coefficients of a Finite Impulse Response (FIR) filter, often known as TF filter (TFF) [100]. If $K_1 \geq 2$, the estimates of the synchrophasor, frequency and ROCOF in t_r can be obtained from the first three components of $\hat{\mathbf{p}}(t_r)$, i.e., $\{\hat{X}_1^{(k)}\}_{k \in \{0,1,2\}}$.

In fact, $\hat{X}_1^{(0)}$, as mentioned above, is the estimate of the phasor itself, while the frequency can be obtained as

$$\hat{f}_1(t_r) = \frac{1}{2\pi} \left(\omega_0 + \frac{\Im \left[\hat{X}_1^{(1)} \cdot \hat{X}_1^{(0)*} \right]}{\left| \hat{X}_1^{(0)} \right|^2} \right) \quad (2.27)$$

and, finally, ROCOF can be calculated by a similar formula that also involves the second derivative as:

$$\widehat{\text{ROCOF}} = \frac{\Im \left\{ \hat{X}_1^{(2)} e^{-j\angle \hat{X}_1^{(0)}} \right\}}{2\pi \left| \hat{X}_1^{(0)} \right|} - \frac{\Re \left\{ \hat{X}_1^{(1)} e^{-j\angle \hat{X}_1^{(0)}} \right\} \Im \left\{ \hat{X}_1^{(1)} e^{-j\angle \hat{X}_1^{(0)}} \right\}}{\pi \left| \hat{X}_1^{(0)} \right|^2} \quad (2.28)$$

The quality of the estimates strongly depends on how well the adopted TF model is able to match the actual signal in the considered observation interval. In particular, the estimate is exact if the signal in the observed time window belongs to the subspace spanned by the selected TF model, thus implying that $\mathbf{x}(t_r)$ can be expressed as a linear combination of the columns of $\bar{\mathbf{B}}$.

However, in practical situations, the model is not exact, therefore $\mathbf{x}(t_r)$ is outside the subspace generated by the columns of $\bar{\mathbf{B}}$. Under these conditions, it is beneficial to favor a smaller residual in the neighborhood of the measurement instant t_r , at the expense of a worse fitting near the edges of the samples record. For this purpose, it is possible to introduce a weighting vector $\mathbf{w}$ (see [101], [114] for TFFs, and [95] for TFM, which is better discussed in what follows) while minimizing the Euclidean norm of the weighted residual $\mathbf{r}_w(t_r) = \mathbf{W}\mathbf{r}(t_r)$, with $\mathbf{W} = \text{diag}(\mathbf{w})$. From a mathematical point of view, it corresponds to a weighted LS (WLS) problem, i.e., a minimization of the residuals norm, that can be written as

$$\hat{\mathbf{p}}(t_r) = \arg \min_{\bar{\mathbf{p}}} \left\| \bar{\mathbf{B}}_w \bar{\mathbf{p}} - \mathbf{x}_w(t_r) \right\| \quad (2.29)$$

where

$$\bar{\mathbf{B}}_w = \mathbf{W}\bar{\mathbf{B}} \quad \mathbf{x}_w(t_r) = \mathbf{W}\mathbf{x}(t_r). \quad (2.30)$$

The solution of (2.29) results as

$$\hat{\mathbf{p}}(t_r) = \bar{\mathbf{B}}_w^\dagger \mathbf{x}_w(t_r) = \bar{\mathbf{H}}_w \mathbf{x}(t_r) \quad (2.31)$$

with $\bar{\mathbf{H}}_w$ defining a new bank of TFFs, generally having wider passbands because of the introduction of weighting.

The above described TFFs can be also designed starting from a different reference frequency $\omega \neq \omega_1$ using seamlessly (2.25), (2.26) and $\mathbf{w}$ to build a new $\bar{\mathbf{B}}_w(\omega)$ and, as a consequence, a new $\bar{\mathbf{H}}_w(\omega)$ to solve the estimation problem. This results in a frequency shift of both passband and zeros of TFFs.

When looking at (2.2), the presence of significant narrowband components

(such as harmonics) in the term $d(t)$ generates mismatch between signal and TF model. Under these conditions, the WLS solution (2.31) is not able to effectively separate the signal from the disturbance subspace; the result is that the obtained estimates are affected by the presence of $d(t)$, thus accuracy is degraded.

A possible solution to overcome this problem is enlarging the signal model subspace to include the most relevant narrowband components, whose spectral content is located around each frequency in the set $\{\omega_q\}_{q \in \{1, \dots, Q\}}$. Their interference on fundamental synchrophasor, frequency, and ROCOF estimates is thus dramatically mitigated. From an implementation point of view, thanks to the narrowband assumption, the signal contribution coming from the generic q -th disturbance can be written through a Taylor expansion of the corresponding synchrophasor around t_r , truncated to the K_q -th degree. Like Eq. (2.21), a general expression for a generic sampled value $x(t_r + nT_s)$ that incorporates multiple narrowband contributions can now be written as:

$$x(t_r + nT_s) = \sum_{q=1}^Q \sum_{k=0}^{K_q} \frac{(nT_s)^k}{k!} \left[\frac{\bar{X}_q^{(k)}}{\sqrt{2}} e^{j\omega_q(t_r + nT_s)} + \frac{X_q^{(k)}}{\sqrt{2}} e^{-j\omega_q(t_r + nT_s)} \right] + r \quad (2.32)$$

where $\bar{X}_q^{(k)}$ is the k -th order derivative of the synchrophasor for the q -th narrowband component. In Eq. (2.32) and also in the following, to abbreviate the formulas, $\bar{X}_q^{(k)}$ does not report the explicit term (t_r) as in Eq. (2.21) and (2.23), but anyway it refers to the derivative calculated in t_r (the same considerations apply to the residual r). This leads to the so-called TFM models [98], which can be expressed in the form (2.22), but now

$$\bar{\mathbf{p}} = \left[\bar{\mathbf{p}}_1^\top \bar{\mathbf{p}}_2^\top \cdots \bar{\mathbf{p}}_Q^\top \right]^\top \quad (2.33)$$

with

$$\begin{aligned} \bar{\mathbf{p}}_q &= \left[\bar{X}_q^{(0)} \cdots \bar{X}_q^{(K_q)} \quad X_q^{(0)} \cdots X_q^{(K_q)} \right]^\top \\ \bar{\mathbf{B}} &= \left[\bar{\Phi}(\omega_1) \mathbf{A}(K_1) \quad \bar{\Phi}(\omega_1) \mathbf{A}(K_1) \cdots \bar{\Phi}(\omega_Q) \mathbf{A}(K_Q) \right]^\top. \end{aligned} \quad (2.34)$$

The spectral components to be embedded in the signal model (namely its spectral support) can be either retrieved through an iterative procedure, or they can be included *ex-ante*. In this respect, significant low-order harmonics are often present in power system waveforms, and they may have noticeable impact on the estimates. For this reason, the adopted TFM model often considers components whose frequencies are the first multiples of f_0 .

2.5 Left and right Taylor–Fourier estimates

PMU algorithms based on TFM models assume that the time evolutions of the magnitude and phase angle of the fundamental term are smooth. Under this condition, given the typical length of the observation window (bounded by the maximum latency constraint), the fundamental component can be accurately tracked with K_1 typically equal to 2 or 3 (see, for instance, [100], [105], [115]). As mentioned above, $K_1 \geq 2$ is needed to compute the first and second order derivatives of the phase angle. Higher expansion degrees are not generally adopted: on the one hand, measurement bandwidth is increased, but on the other hand estimates become more prone to spectral interference from both narrowband components and wideband noise. In some cases, this drawback can outweigh the potential advantage.

During faults, switching or in case of sudden load variations, voltage and current waveforms may exhibit an abrupt transition between two different steady-state conditions, which could be idealized as a step-like change in magnitude or phase of the underlying phasor. In order to assess performance in the presence of such events, as discussed in Section 2.2 the PMU Std prescribes compliance tests where the applied sinusoidal signal at nominal frequency contains jumps either in amplitude or in phase. Assuming that a step occurs in the time instant t_{step} , when $t_r < t_{\text{step}}$, an ideal PMU algorithm should return the synchrophasor value corresponding to the steady-state before the step, zero frequency deviation (with respect to the nominal value) and zero ROCOF; when $t_r > t_{\text{step}}$, the measured synchrophasor should switch to the steady-state value after the step change, while frequency deviation and ROCOF should remain zero.

In general, this does not happen with real-world PMU algorithms: the relationship between input samples and measurements is dynamic, hence it unavoidably slows down the transition between the two steady-state conditions. For example, in case of PMU methods based on FIR filtering, the duration of the transition corresponds to the length of the filter. The response time quantifies this phenomenon, since it is the duration of the time interval across the step event where steady-state accuracy limits for synchrophasor, frequency and ROCOF measurements are violated.

It is immediately evident that a TFM model of the observed samples does not represent an effective tool to deal with step changes, just as a Taylor expansion struggles in approximating discontinuities. As previously pointed out, improving responsiveness by enlarging the bandwidth of the TFFs (namely increasing the Taylor expansion degree or adopting proper weights) is not a good choice. Therefore, different approaches should be developed to enhance performance of TFM-based PMU algorithms during step variations.

From a different point of view, the reporting instant splits the $2N + 1$ sample window into two $N + 1$ sample sub-windows: one located at the left, the other at the right of t_r (both of them include the sample in t_r). In the presence of a step within the full window, if $t_r < t_{\text{step}}$ the left sub-window does not contain the step, and all its samples are related to the steady-state condition before the step. Conversely, when $t_r > t_{\text{step}}$, the right sub-window does not include the step event, while the corresponding samples are tied to the steady-state after the step. Therefore, if one would be able to perform measurements using just the left sub-window for $t_r < t_{\text{step}}$, while switching to the right sub-window as t_r exceeds t_{step} , in principle it would be possible to approach the ideal behavior in the presence of step changes.

The starting point of the idea is having available a method that enables obtaining the measurements in t_r by processing just the samples in either the left or right sub-window. A possible solution is adopting the TFM approach, but considering asymmetric expansions of the embedded components, looking exclusively at the left or right of t_r . The left and right sub-windows $\mathbf{x}_L(t_r)$ and $\mathbf{x}_R(t_r)$, corresponding to the last and first $N + 1$ components of $\mathbf{x}_w(t_r)$, respectively, can be defined. Left and right TFM models can be written as

$$\mathbf{x}_S(t_r) = \bar{\mathbf{B}}_S \bar{\mathbf{p}}_S(t_r) + \mathbf{r}_S(t_r) \quad (2.35)$$

with $S \in \{L, R\}$ while $\bar{\mathbf{B}}_L$ and $\bar{\mathbf{B}}_R$ correspond to the last and first $N + 1$ rows of $\bar{\mathbf{B}}$. $\bar{\mathbf{p}}_L(t_r)$ and $\bar{\mathbf{p}}_R(t_r)$ represent the synchrophasors and their derivatives in the left and right sub-window, respectively.

The weighting vectors $\mathbf{w}_L$ and $\mathbf{w}_R$ (and the corresponding diagonal matrices $\mathbf{W}_S = \text{diag}(\mathbf{w}_S)$) made of the last and first $N + 1$ components of $\mathbf{w}$ can be also introduced. Under the usual assumptions, the estimates $\hat{\mathbf{p}}_L(t_r)$ and $\hat{\mathbf{p}}_R(t_r)$ of the synchrophasors and their derivatives in t_r , which are computed, respectively, from the left and right portion of the samples (that will be often referred to as left and right estimates), are obtained with the WLS approach, hence minimizing the Euclidean norm of $\mathbf{r}_{w,S}(t_r) = \mathbf{W}_S \mathbf{r}_S(t_r)$. Their general expression is given by

$$\hat{\mathbf{p}}_S(t_r) = \bar{\mathbf{B}}_{w,S}^\dagger \mathbf{x}_{w,S}(t_r) = \bar{\mathbf{H}}_{w,S} \mathbf{x}_S(t_r) \quad (2.36)$$

with $\bar{\mathbf{B}}_{w,S} = \mathbf{W}_S \bar{\mathbf{B}}_S$ and $\mathbf{x}_{w,S}(t_r) = \mathbf{W}_S \mathbf{x}_S(t_r)$.

From (2.36), two different banks of filters, represented by the rows of $\bar{\mathbf{H}}_{w,L}$ and $\bar{\mathbf{H}}_{w,R}$, for left and right estimation, respectively, are found. It is interesting to notice that $\bar{\mathbf{H}}_{w,L}$ corresponds to causal filters, while $\bar{\mathbf{H}}_{w,R}$ defines purely non-causal ones, with respect to t_r . Through this complementarity, it is guaranteed that one of the two sets of estimates (phasor, frequency and ROCOF) is free from artifacts due to the step.

2.5.1 Criteria for selecting the Taylor-Fourier expansion

The next problem to be faced is how to exploit the estimates computed in the left and right sub-windows in order to achieve exemplary responsiveness in the presence of fast dynamics, without sacrificing disturbance rejection under slowly-varying conditions.

In [116], the proposed PMU algorithm computes the estimates obtained from three TFM models: one covering the full observation window, the second considering the left sub-window, the third based on the right sub-window.

According to the approach introduced in Section 2.5, the algorithm should return the left estimates if a step is present in the right sub-window, or the right estimates if there is a step in the left sub-window. Since both the left and right TFM models are exact under steady-state conditions at rated frequency, an ideal step behavior can be reached in the presence of a rule that selects the correct estimate.

Left and right estimates are performed on a relatively short time window; moreover, they do not make full use of the samples in the neighborhood of the reporting instant that usually carry the most relevant information. Because of this, if the underlying synchrophasor of the observed signal has a smooth time evolution, measurements obtained from the full window are more accurate and exhibit better robustness with respect to both narrowband and wideband disturbances. Therefore, it is desirable that these measurements are returned thanks to a proper rule detecting such operating condition.

Steps can be classified as abrupt transients that do not belong to the TFM model subspace and that occur in either the left or the right sub-window. The most obvious consequence of a step in one of the sub-windows is that the corresponding misfit increases noticeably with respect to the other. Moreover, the larger of the two misfits is significantly higher than those occurring under regular operation. Therefore, it is possible to detect the step when

$$\begin{cases} \max(\|\bar{\mathbf{B}}_L \hat{\mathbf{p}}_L - \mathbf{x}_L\|, \|\bar{\mathbf{B}}_R \hat{\mathbf{p}}_R - \mathbf{x}_R\|) > \sigma_{th} \\ \frac{\max(\|\bar{\mathbf{B}}_L \hat{\mathbf{p}}_L - \mathbf{x}_L\|, \|\bar{\mathbf{B}}_R \hat{\mathbf{p}}_R - \mathbf{x}_R\|)}{\min(\|\bar{\mathbf{B}}_L \hat{\mathbf{p}}_L - \mathbf{x}_L\|, \|\bar{\mathbf{B}}_R \hat{\mathbf{p}}_R - \mathbf{x}_R\|)} > \rho_{th} \end{cases} \quad (2.37)$$

where σ_{th} and ρ_{th} are predetermined threshold values. It is worth noting that if the first condition is not taken into account, there is the serious risk that a step is erroneously detected since it may happen that both misfits are very small in magnitude, but their ratio exceeds ρ_{th} .

If (2.37) is false, it means that probably the samples do not contain an abrupt transient, hence the estimates performed on the full window are returned. On the contrary, if (2.37) is true, estimates on the sub-window having the lowest misfit are selected.

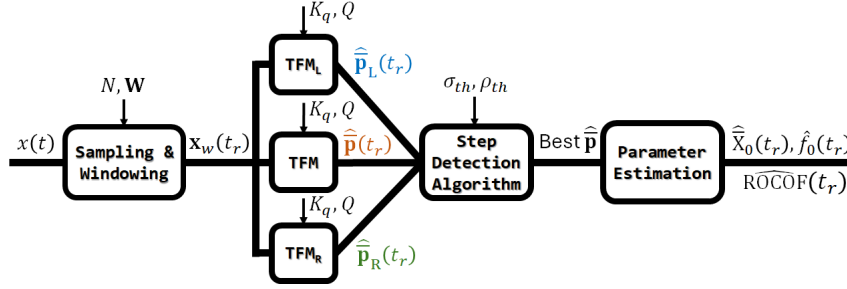


Figure 2.3: Flow chart of the proposed TFM_{LR} algorithm.

Fig. 2.3 shows the block diagram of the algorithm called TFM_{LR} (TFM with left and right estimates). To summarize, the algorithm first samples and applies the vector $\mathbf{w}$ to $\mathbf{x}(t_r)$ obtaining $\mathbf{x}_w(t_r)$. After that, TFM is carried out with the entire window and the left and right half-windows in parallel obtaining $\hat{\mathbf{p}}(t_r)$, $\hat{\mathbf{p}}_L(t_r)$ and $\hat{\mathbf{p}}_R(t_r)$, respectively. These estimates are passed to the step detection algorithm, which chooses the one with the best quality using the criteria of Eq. (2.37). Finally, the synchrophasor, frequency, and ROCOF are estimated from the chosen synchrophasor derivative estimates, using Eq. (2.27) and (2.28). For the sake of clarity, in the following graphs, the estimate that refers to the left or right sub-window will be represented in blue and green, respectively.

2.5.2 Blending left and right estimates

As will be discussed later in Section 2.6, the above described technique has proven to be extremely effective, but such performance is attained thanks to the careful tuning of the threshold values σ_{th} and ρ_{th} . Such thresholds depend on several factors, including the structure of the adopted TFM model, the noise and disturbance level, the minimum amplitude step that has to be detected. In order to overcome this limitation, a method to blend the left and right estimates into a unique value not affected by user-selected thresholds has been proposed in [117]. The basic ideas are:

- The returned estimate should rely more on $\hat{\mathbf{p}}_L$ or $\hat{\mathbf{p}}_R$ according to their relative quality level;
- If the quality of $\hat{\mathbf{p}}_L$ and $\hat{\mathbf{p}}_R$ is virtually identical, the estimate should correspond to that obtained from a TFM model fitted on the whole set of samples;
- The transition between left, full and right estimate should be seamless.

CHAPTER 2. ALGORITHMS FOR SYNCHROPHASOR ESTIMATION IN
PRESENCE OF HIGH DYNAMICS AND TRANSIENTS

In order to reach this target, it is assumed to have a proper TFM model and a weighing vector $\mathbf{w}$, typically generated from a window function. For a given vector of samples, the WLS estimates of the synchrophasors and their derivatives (that is $\hat{\mathbf{p}}$) is obtained through (2.31), but adopting a new, dynamically-adjusted vector of weights $\mathbf{w}_{\text{LR}} = \mathbf{w}_\alpha \circ \mathbf{w}$, where $\circ$ denotes the Hadamard product, while

$$\mathbf{w}_\alpha = \begin{bmatrix} \alpha_R \mathbf{1}_N \\ 1 \\ \alpha_L \mathbf{1}_N \end{bmatrix} \quad (2.38)$$

having introduced $\mathbf{1}_N$ as the size N column vector of ones. The coefficients α_L and α_R appearing in (2.38) are

$$\begin{aligned} \alpha_L &= \min(1 - \lambda, 1) \\ \alpha_R &= \min(1 + \lambda, 1). \end{aligned} \quad (2.39)$$

Both α_L and α_R depend on a real-valued parameter $\lambda \in [-1, 1]$, which should reflect the relative quality of left and right estimates. In this respect, $\lambda = -1$ means that it is worth trusting exclusively the left estimate; in fact, in this case $\alpha_L = 1$, $\alpha_R = 0$, hence the WLS solution obtained with the weighting vector $\mathbf{w}_{\text{LR}}$ corresponds to the left estimate given by (2.36) with $S = \text{L}$. The opposite situation should result in $\lambda = 1$: it implies $\alpha_L = 0$, $\alpha_R = 1$, hence $\hat{\mathbf{p}} = \hat{\mathbf{p}}_R$. Finally, if $\lambda = 0$ it results in $\alpha_L = \alpha_R = 1$, so that $\mathbf{w}_{\text{LR}} = \mathbf{w}$: in that case, the usual WLS estimate computed from the whole set of samples is obtained.

Having explained the foundation of the method, it is interesting to derive the analytic expression of $\hat{\mathbf{p}}$, thus investigating its connection to $\hat{\mathbf{p}}_L$ and $\hat{\mathbf{p}}_R$. Reminding the WLS solution (2.31) with the newly defined weighting vector $\mathbf{w}_{\text{LR}}$, introducing H as the conjugate transpose operator, it requires computing

$$\hat{\mathbf{p}}(t_r) = \bar{\mathbf{H}}_{w,\text{LR}} \mathbf{x}(t_r) = (\bar{\mathbf{B}}_w^H \mathbf{W}_\alpha^2 \bar{\mathbf{B}}_w)^{-1} \bar{\mathbf{B}}_w^H \mathbf{W}_\alpha^2 \mathbf{x}_w(t_r) \quad (2.40)$$

where $\mathbf{W}_\alpha = \text{diag}(\mathbf{w}_\alpha)$, while $\bar{\mathbf{B}}_w$ and $\mathbf{x}_w(t_r)$ are still given by (2.30). $\bar{\mathbf{H}}_{w,\text{LR}}$ represents a bank of FIR filters, whose behavior can be modified on the run by acting on the parameter λ . $\bar{\mathbf{B}}_w$ and $\mathbf{x}_w(t_r)$ can be partitioned as

$$\bar{\mathbf{B}}_w = \begin{bmatrix} \bar{\mathbf{B}}_{w,\text{R-C}} \\ \bar{\mathbf{b}}_{w,\text{C}}^T \\ \bar{\mathbf{B}}_{w,\text{L-C}} \end{bmatrix} \quad \mathbf{x}_w(t_r) = \begin{bmatrix} \mathbf{x}_{w,\text{R-C}}(t_r) \\ x_{w,\text{C}}(t_r) \\ \mathbf{x}_{w,\text{L-C}}(t_r) \end{bmatrix}. \quad (2.41)$$

$\bar{\mathbf{b}}_{w,\text{C}}^T = \bar{\mathbf{B}}_w[N+1, *]$ is the central row of $\bar{\mathbf{B}}_w$, while $\bar{\mathbf{B}}_{w,\text{R-C}}$ and $\bar{\mathbf{B}}_{w,\text{L-C}}$ contain its first and last N rows, respectively. Similarly, $x_{w,\text{C}}(t_r)$ is the central element of $\mathbf{x}_w(t_r)$, while $\mathbf{x}_{w,\text{R-C}}(t_r)$ and $\mathbf{x}_{w,\text{L-C}}(t_r)$ are made of the first and last N elements. Substituting (2.41) into (2.40) allows obtaining

$$\begin{aligned} \hat{\mathbf{p}}(t_r) &= (\alpha_L^2 \bar{\mathbf{G}}_{w,\text{L-C}} + \bar{\mathbf{G}}_{w,\text{C}} + \alpha_R^2 \bar{\mathbf{G}}_{w,\text{R-C}})^{-1} \cdot \\ &\quad (\alpha_L^2 \bar{\mathbf{B}}_{w,\text{L-C}}^H \mathbf{x}_{w,\text{L-C}} + \bar{\mathbf{b}}_{w,\text{C}} x_{w,\text{C}} + \alpha_R^2 \bar{\mathbf{B}}_{w,\text{R-C}}^H \mathbf{x}_{w,\text{R-C}}) \end{aligned} \quad (2.42)$$

having defined the matrices $\bar{\mathbf{G}}_{w,S-C} = \bar{\mathbf{B}}_{w,S-C}^H \bar{\mathbf{B}}_{w,S-C}$ and $\bar{\mathbf{G}}_{w,C} = \mathbf{b}_{w,C} \bar{\mathbf{b}}_{w,C}^T$.

It is worth noting that $\bar{\mathbf{B}}_{w,R-C}$ and $\bar{\mathbf{B}}_{w,L-C}$ can be respectively obtained by removing the last row from $\bar{\mathbf{B}}_{w,R}$ and the first row from $\bar{\mathbf{B}}_{w,L}$, as defined in Section 2.5. Manipulating (2.42) while using (2.36), it can be obtained

$$\begin{aligned} \hat{\mathbf{p}}(t_r) &= (\alpha_L^2 \bar{\mathbf{G}}_{w,L} + (1 - \alpha_L^2 - \alpha_R^2) \bar{\mathbf{G}}_{w,C} + \alpha_R^2 \bar{\mathbf{G}}_{w,R})^{-1} \cdot \\ &(\alpha_L^2 \bar{\mathbf{G}}_{w,L} \hat{\mathbf{p}}_L + (1 - \alpha_L^2 - \alpha_R^2) \mathbf{b}_{w,C} x_{w,C} + \alpha_R^2 \bar{\mathbf{G}}_{w,R} \hat{\mathbf{p}}_R) \end{aligned} \quad (2.43)$$

with $\bar{\mathbf{G}}_{w,S} = \bar{\mathbf{B}}_{w,S}^H \bar{\mathbf{B}}_{w,S}$. According to the previous equation, $\hat{\mathbf{p}}(t_r)$ can be computed as a linear combination of the estimates obtained from the left and right sub-window. A correction related with the central sample $x_{w,C}$ is also present, since it corresponds to the overlap between $\mathbf{x}_{w,L}(t_r)$ and $\mathbf{x}_{w,R}(t_r)$ that have been processed to obtain $\hat{\mathbf{p}}_L$ and $\hat{\mathbf{p}}_R$.

The last aspect to be discussed is the definition of the coefficient λ , which should affect the balance between left and right estimates on the basis of their relative trustworthiness. In particular, it should dramatically reduce the weight of a sub-window as long as it is likely to include an abrupt transient. Such a fast transition is clearly well outside the TFM model subspace: this results in a noticeable increase in the Euclidean norm of the corresponding weighted residual, which is the cost function minimized by the WLS problem. $\|\mathbf{r}_{w,L}\|$ and $\|\mathbf{r}_{w,R}\|$ could be thus considered as inversely proportional to the quality level of the corresponding estimates. It is therefore reasonable to choose α_L and α_R , namely the scaling of the weights of the samples in the left and right sub-window, so that they are inversely proportional to $\|\mathbf{r}_{w,L}\|$ and $\|\mathbf{r}_{w,R}\|$. According to this choice, using (2.39) it can be derived

$$\lambda = \begin{cases} -1 + \frac{\|\mathbf{r}_{w,L}\|}{\|\mathbf{r}_{w,R}\|} & \text{if } \|\mathbf{r}_{w,R}\| \geq \|\mathbf{r}_{w,L}\| \\ 1 - \frac{\|\mathbf{r}_{w,R}\|}{\|\mathbf{r}_{w,L}\|} & \text{if } \|\mathbf{r}_{w,R}\| < \|\mathbf{r}_{w,L}\|. \end{cases} \quad (2.44)$$

Summarizing, once the TFM model (thus matrix $\bar{\mathbf{B}}$) and the weighting vector $\mathbf{w}$ are defined, the steps required to obtain synchrophasor, frequency and ROCOF estimates from $\mathbf{x}(t_r)$ by adopting the proposed method, called TFM_{WRLR} (TFM with weighted reconstruction from left and right estimates), are the following:

1. Apply $\mathbf{w}$ to $\mathbf{x}(t_r)$, obtain $\hat{\mathbf{p}}_L(t_r)$, $\hat{\mathbf{p}}_R(t_r)$ with (2.36);
2. Evaluate the norms of the left and right residuals, which enable obtaining λ (i.e., also α_L and α_R) and thus computing $\hat{\mathbf{p}}(t_r)$ from (2.43);
3. Synchrophasor, frequency and ROCOF estimates in the measurement instant t_r are finally derived from the first three components of $\hat{\mathbf{p}}(t_r)$ (see Eq. (2.27) and (2.28)).

These stages are also schematically summarized by means of the flow chart shown in the Fig. 2.4. For the sake of clarity, the recombined estimate $\hat{\mathbf{p}}$ is

shown in yellow, while quantities referring to the left or right sub-window, as in Fig. 2.3, are represented in blue and green, respectively. From now on these colors will be used in the following graphs.

One important novelty of the proposed method is that it does no longer rely on thresholds to choose between left and right estimates. Now a merging procedure based on weighting is adopted, without a dedicated step detection stage. A minimum residual threshold might be used to address clean signals as in [116] (if $\|\mathbf{r}_{w,R}\|$ and $\|\mathbf{r}_{w,L}\|$ are very small, λ is ill-defined) but this becomes irrelevant as soon as wideband noise, even very low, is present.

2.6 Tests and results

This section will gradually present the results achieved in the steady-state and dynamic tests firstly by the proposed method TFM_{LR} and immediately after those obtained by the proposed method TFM_{WRLR} .

2.6.1 Proposed methods configuration

The method called TFM_{LR} has been implemented in MATLAB environment considering $f_0 = 50$ Hz, adopting $T_s = 10^{-4}$ s (sampling rate $f_s = 10$ kHz) and $N_w = 781$, which guarantees compliance with P-class latency requirement in PMU Std (2 nominal cycles at 50 frames per second, fps), leaving enough time for synchrophasor computation. A third order expansion has been adopted for the fundamental term, hence $K_0 = 3$. Harmonics up to the fourth order have been included in the TFM model, using first order expansions (i.e., $K_q = 1$, $q \in \{1, 2, 3\}$), and thus higher order harmonics are not part of it regardless their possible presence in the signals. Thresholds ρ_{th} and σ_{th} have been set, respectively, to 10 and $5 \cdot 10^{-4}$ times the estimated RMS value of the fundamental $\hat{X}_1$ computed with the TFM on the entire window. For this model, the rectangular window was employed, i.e., $\mathbf{w}$ is a vector of all ones, thus the samples are not weighted differently from each other.

In order to validate the proposed TFM_{LR} , its performance is characterized over the entire test-set for P-class compliance prescribed by the PMU Std.

Passing to the second proposed method, the TFM_{WRLR} was also implemented in MATLAB environment considering again $f_0 = 50$ Hz, but adopting a longer window $T_w = 180 \text{ ms} + T_s$ to meet M-class requirements. This corresponds to a window of about 9 nominal cycles (and thus left and right half windows of about 4.5 cycles), while keeping an odd number of samples, as TFM_{LR} and according to the assumption of Sections 2.4 and 2.5.

The target RR for both the proposed algorithms is 50 fps and this value is

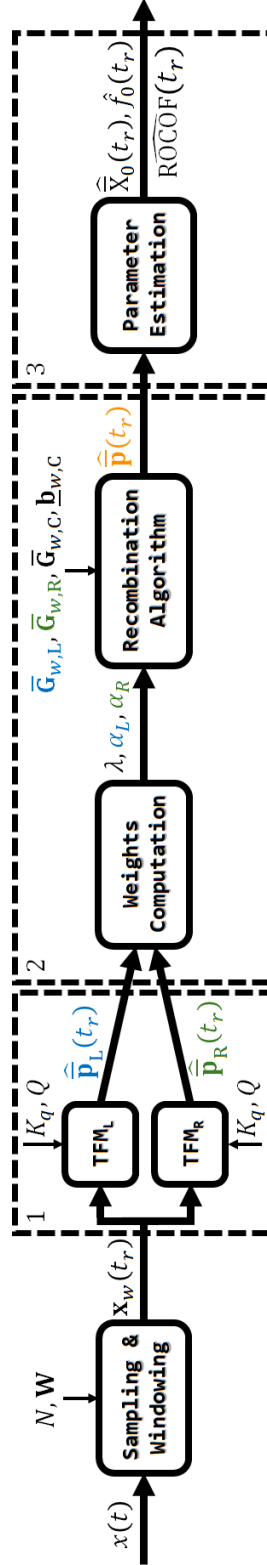


Figure 2.4: Flow chart of the proposed TFM_{WRLR} algorithm.

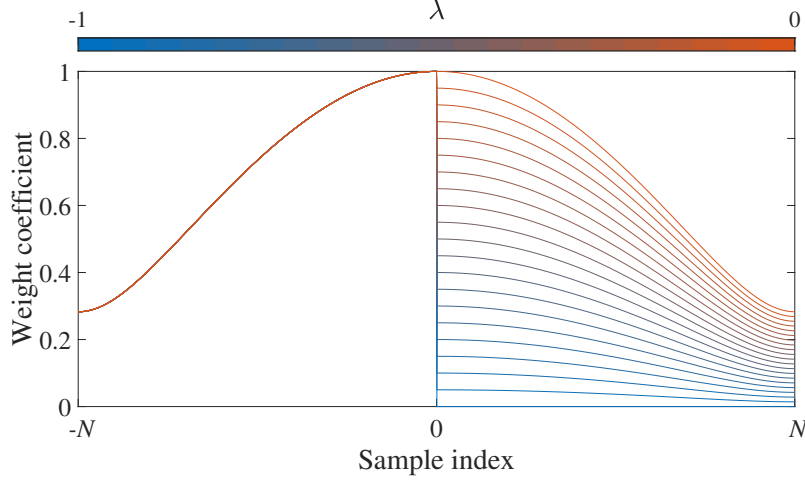


Figure 2.5: Weight coefficients (components of $\mathbf{w}_{\text{LR}}$) parametrized by $-1 \leq \lambda \leq 0$.

employed for setting the limits in terms of PMU pass-bandwidth and standard requirements. The degree expansions of the various components and the harmonic order of the TFM_{WRLR} have been maintained the same as TFM_{LR} (i.e., $K_1 = 3$, $K_q = 1$, $q \in \{2, 3, 4\}$), and thus higher order harmonics are not part of it regardless their possible presence in the signals. The reason for this choice for both the models is that, thanks to the frequency separation and the decaying sidelobes of the filters, they produce weak interference with the fundamental estimates.

To largely filter out harmonics and interharmonics that have not been added to the TFM model, the choice of window is important. For the TFM_{WRLR} , vector $\mathbf{w}$ has been generated as the square root of the Hamming window. This is a difference between the two methods presented and derives from the need in TFM_{WRLR} to be more robust to interference with respect to TFM_{LR} due to the different target performance class. Fig. 2.5 shows, for $-1 \leq \lambda \leq 0$, the resulting values of the weight coefficients applied to the samples, namely the elements of vector $\mathbf{w}_{\text{LR}}$. Opposite (positive) values of λ would have resulted in mirrored weights with respect to sample index $n = 0$.

Having defined the TFM model, for a given value of λ it is possible to compute the frequency responses of the FIR filters defined by the rows of matrix $\bar{\mathbf{H}}_{w,\text{LR}}$. In this respect, Fig. 2.6 shows the magnitude response of the filter that allows extracting the synchrophasor for $-1 \leq \lambda \leq 0$ (the shape is not affected by the sign of λ). Amplitude is unitary at the reference fundamental frequency, while multiple zeros are present at the harmonic frequencies included into the

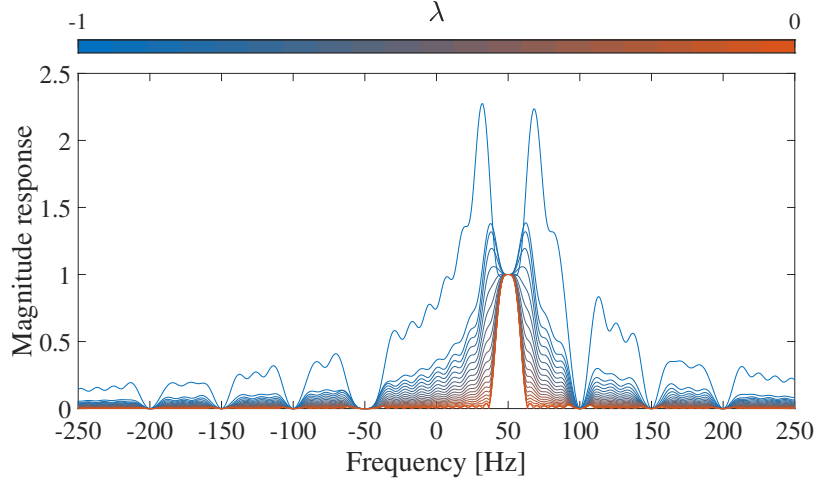


Figure 2.6: Magnitude response of the family of filters that enable estimating $\bar{X}_1$ parametrized by $-1 \leq \lambda \leq 0$.

model. It is worth highlighting that as $|\lambda|$ increases, a significant rise of the sidelobes occurs. This results in higher sensitivity to wideband disturbances, hence it has key importance that λ is close to zero, except in the presence of abrupt transients.

The reference frequency of the TFM_{WRLR} model is updated according to the previous frequency measurement using a coarse 1 Hz grid to address large frequency deviations as in [96], [105]. This means that when frequency undergoes strong variations with respect to its rated value, like in off-nominal frequency and ramp tests, a few TFM model transitions can occur. Being them based on pre-computed filters, they do not affect computational burden.

To reduce the slight fluctuations that can occur in noisy conditions, the value of $|\lambda|$ can be rounded to 1, e.g. when $|\lambda| > 0.86$ (i.e., when a 2% ratio among left and right, or vice-versa, residual energies is obtained) at run time.

It is important to highlight that the specific methods configuration is mainly considered here as an example of settings that allow the solution found using the TFM models on the full window to be compliant with P-class and M-class requirements of the PMU Std for TFM_{LR} and TFM_{WRLR} , respectively. The parameters (expansion degrees, window length, included components, weighing, etc.) arise from a trade-off between different needs (harmonics or interharmonic rejection, dynamics tracking, etc.) and thus different settings allow compliance.

Moreover, the proposed approaches can be also combined with other techniques, in particular those based on an adaptive TFM model for canceling narrowband interferers [98], [95]: in this case, M-class compliance can be achieved

with lower latency. However, these solutions have not been considered here for the sake of a more concise explanation. In this respect, there is no presumption here to reach “optimal” performance, but rather to clearly illustrate the potentialities of the proposal based on merging left and right TF expansions. From a different point of view, it is very significant to implement the method starting from a rather long TFM model, as the one chosen for TFM_{WRLR} , which typically implies slow responsiveness.

2.6.2 Steady-state and dynamic tests

In order to account for the measurement noise introduced by the acquisition process, the test waveforms are corrupted by additive white uniform noise with a Signal-to-Noise Ratio (SNR) of 80 dB. Results are compared with those achieved by the conventional TFM algorithm (TFM from here on) with the same full window length, modeled components and expansion orders.

If not otherwise specified, the test waveforms have 10 s overall duration and estimates have been computed sample by sample. This measurement rate has been used only for testing purposes since, for each condition, this yields a statistically relevant sample of phasor, frequency and ROCOF measurements. Results are thus more representative and define a conservative scenario in terms of errors. Indeed, this allows considering also the dependence on different combinations of signal parameters (i.e., frequency and initial phase). Nevertheless, it is important to notice that the two algorithms are intended to operate with a RR of 50 fps, and this value is adopted when looking at the PMU Std limits. For step tests, in particular, the sample-by-sample approach allows a better excitation of the algorithm and a finer description of the step response with respect to the usual PMU testing procedure.

For each test, performance in terms of TVE and absolute errors of AE, PE, FE, and RFE were assessed. In this regard, Table 2.3 reports the worst-case errors of the TFM_{LR} in all the static and dynamic compliance tests and, when applicable, it compares them against the corresponding PMU Std requirements. In all the test conditions, the proposed method proves to be fully P-class compliant.

In the static tests, TFM_{LR} provides the exact same results as TFM. This proves the validity and robustness of the detection criteria, and of the selection of σ_{th} and ρ_{th} . Indeed, the expansion is computed on the entire window. The worst performance is obtained in the harmonic distortion test with Total Harmonic Distortion (THD) of 1 %, specifically in the presence of the 5th order harmonic. This result descends directly from the structure of the TFM model, which only accounts for the first four harmonics. The uncompensated effect of the 5th

Table 2.3: Worst-Case Errors for P-Class Steady-State and Dynamic Compliance Tests (TFM and TFM_{LR} algorithms)

Test	AE [%]	PE [crad]	TVE [%]	Std [%]	FE [mHz]	Std [mHz]	RFE [Hz/s]	Std [Hz/s]
nominal	$2.23 \cdot 10^{-3}$	$1.88 \cdot 10^{-3}$	0.003	1	0.295	5	0.014	0.4
signal freq. ($f_0 \pm 2$ Hz)	$2.02 \cdot 10^{-2}$	$3.12 \cdot 10^{-3}$	0.021	1	1.165	5	0.046	0.4
harm. dist. (THD = 1 %)*	$2.89 \cdot 10^{-2}$	$2.68 \cdot 10^{-2}$	0.034	1	3.728	5	0.377	0.4
ampl. mod. ($f_m \leq 2$ Hz)	$4.89 \cdot 10^{-3}$	$2.62 \cdot 10^{-3}$	0.005	3	0.413	60	0.016	2.3
phase mod. ($f_m \leq 2$ Hz)	$3.41 \cdot 10^{-3}$	$3.47 \cdot 10^{-3}$	0.004	3	0.391	60	0.050	2.3
freq. ramp ($R_f = +1$ Hz/s)	$1.67 \cdot 10^{-2}$	$5.28 \cdot 10^{-3}$	0.017	1	0.722	10	-0.111	0.4
freq. ramp ($R_f = -1$ Hz/s)	$1.64 \cdot 10^{-2}$	$5.01 \cdot 10^{-3}$	0.017	1	0.782	10	0.118	0.4

* The worst-case performance is obtained in the presence of the 5th order harmonic.

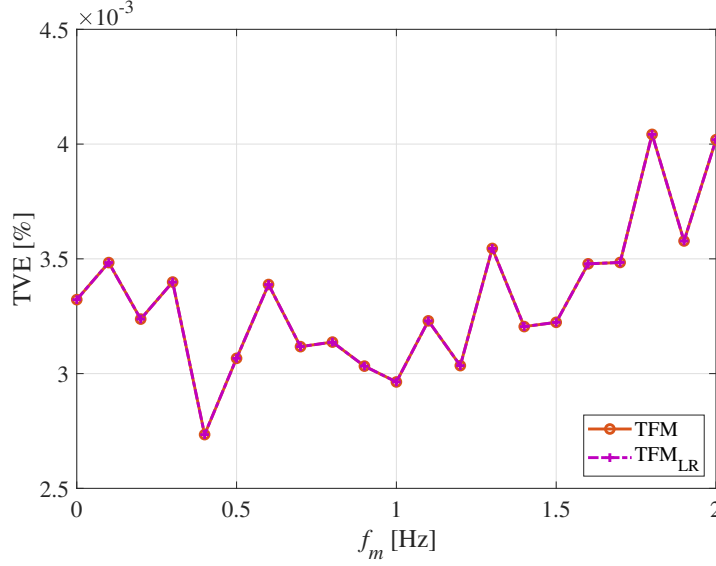


Figure 2.7: Phase modulation test: TVE as function of modulation frequency f_m as produced by TFM and TFM_{LR} in red and purple, respectively.

order harmonic causes a non-negligible error, particularly in terms of FE and RFE, but still within P-class requirements. Similar considerations apply for the signal frequency test: the worst performance is obtained when the fundamental frequency is equal to 52 Hz, due to the non perfectly flat passband response of the obtained TFF.

Considering dynamic conditions, modulation tests exhibit remarkable results. Also in this case, the TFM_{LR} approach performs the expansion on the full window, thus achieving the same errors as the TFM method, which outperforms the corresponding requirements by at least two orders of magnitude. This is clearly visible from Fig. 2.7 that shows the maximum TVE as function of the modulation frequency in the PM test. The plot compares the estimates provided by TFM in red, against the proposed TFM_{LR} in purple. It is worth reminding that the observation window length and expansion orders are exactly the same, the only difference consists in the possibility of splitting the TFM expansion on the left or right side, if needed.

In the frequency ramp tests, the waveforms have a 4-s duration. This allows spanning the whole frequency range of interest (from 48 to 52 Hz) with a constant ROCOF of +1 or -1 Hz/s. The worst performance in terms of FE and RFE is obtained at the beginning of the negative ramp, i.e., with frequency near 52 Hz, consistently with what observed in the static tests. Errors achieved with TFM and TFM_{LR} methods are identical also in this case.

Passing to the second proposal, in order to validate the TFM_{WRLR} method, its performance has been characterized over the entire test-set for M-class com-

pliance prescribed by the PMU Std. For this reason, analogously to Table 2.3, Table 2.4 reports the worst-case values in all the static and dynamic compliance tests for M-class compliance.

Considering all the test conditions, the second proposed method proves to be fully M-class compliant. In the static tests, TFM_{WRLR} , as the first proposed method, also provides virtually the same results as TFM, which are thus not reported in Table 2.4. This proves the validity and robustness of the method. It is important to highlight that, in such conditions, left and right estimates are combined with similar weights and thus it is like computing the expansion on the entire window. Very low errors are achieved both under off-nominal frequency conditions (thanks to the adaptive reference frequency of the TFM model) and in the harmonic distortion test. In that case, again the highest error occurs in the presence of the 5th order harmonic, namely the lowest that has not been embedded in the signal model. Worst performance is obtained in the OOB test with Total Interharmonic Distortion (TIHD) of 10 %. This result derives directly from the selected structure for the TFM model, which only includes the first four harmonics, thus it does not account for possible interharmonics. The uncompensated effect of the interharmonics causes a non-negligible error, particularly in terms of FE and RFE, but it still remains within class M requirements.

Considering dynamic conditions, the TFM_{WRLR} method still reconstructs almost exactly the TFM estimate (i.e., λ is always close to zero), thus achieving the same errors (not reported in Table 2.4). All the maximum errors are largely within the corresponding requirements thanks to the inherently dynamic model of the fundamental component embedded in the TFM expansion. This is particularly effective for modulated signals, while during frequency ramp tests (from 45 to 55 Hz or vice-versa) the frequency tuning also comes into play for keeping the errors low. As previously mentioned, the change of the TFM reference frequency does not impact the computational burden.

The next step is analyzing the impact on the proposed algorithms of different noise levels and sampling rates under nominal conditions. Similar considerations apply to both methods, thus TFM_{LR} results are not reported here for brevity. If the samples are affected by additive, independent noise having variance $\sigma^2 \ll |\bar{X}_1|^2$, given the values of λ and T_w , it is possible to prove that, for M large enough, the estimation errors induced by superimposed noise have Gaussian distributions, whose variances are proportional to σ^2 and inversely proportional to M . Moreover, it is interesting to stress that $\|\mathbf{r}_{w,L}\|^2$ and $\|\mathbf{r}_{w,R}\|^2$ are purely due to noise: therefore, reminding (2.44), λ approaches zero. For this very reason, noise rejection of the TFM_{WRLR} algorithm is expected to be virtually identical to that of the conventional TFM method.

Table 2.4: Worst-Case Errors for M-Class Steady-State and Dynamic Compliance Tests (TFM and TFM_{WRLR} algorithms)

Test	AE [%]	PE [crad]	TVE [%]	Std [%]	FE [mHz]	Std [mHz]	RFE [Hz/s]	Std [Hz/s]
nominal	1.0·10 ⁻³	1.1·10 ⁻³	1.4·10 ⁻³	1	0.07	5	1.6·10 ⁻³	0.1
signal freq. ($f_0 \pm 5$ Hz)	1.9·10 ⁻³	1.7·10 ⁻³	1.9·10 ⁻³	1	0.11	5	2.9·10 ⁻³	0.1
harm. dist. (THD = 10 %)	2.6·10 ⁻³	2.7·10 ⁻³	2.7·10 ⁻³	1	1.89	5	7.9·10 ⁻³	-
OOB (50.0 Hz, TIHD = 10 %)	5.9·10 ⁻²	6.2·10 ⁻²	6.2·10 ⁻²	1.3	9.27	10	0.32	-
OOB (52.5 Hz, TIHD = 10 %)	7.3·10 ⁻²	7.3·10 ⁻²	7.4·10 ⁻²	1.3	8.31	10	0.38	-
OOB (47.5 Hz, TIHD = 10 %)	7.4·10 ⁻²	7.3·10 ⁻²	7.4·10 ⁻²	1.3	8.95	10	0.34	-
ampl. mod. ($f_m \leq 5$ Hz)	0.51	9.6·10 ⁻³	0.51	3	2.34	300	4.7·10 ⁻²	14
phase mod. ($f_m \leq 5$ Hz)	0.23	0.46	0.47	3	23.1	300	4.40	14
freq. ramp ($R_f = +1$ Hz/s)	1.9·10 ⁻³	3.0·10 ⁻³	3.0·10 ⁻³	1	0.09	10	2.5·10 ⁻²	0.2
freq. ramp ($R_f = -1$ Hz/s)	2.0·10 ⁻³	3.3·10 ⁻³	3.3·10 ⁻³	1	0.10	10	2.4·10 ⁻²	0.2

CHAPTER 2. ALGORITHMS FOR SYNCHROPHASOR ESTIMATION IN PRESENCE OF HIGH DYNAMICS AND TRANSIENTS

Table 2.5: Maximum and RMS Errors for Different SNRs in Nominal Conditions (TFM and TFM_{WRLR} algorithms)

SNR [dB]	TVE [%]		FE [mHz]		RFE [Hz/s]	
	Max	RMS	Max	RMS	Max	RMS
60	1.4	$0.6 \cdot 10^{-2}$	7.4	$2.5 \cdot 10^{-1}$	1.7	$0.6 \cdot 10^{-3}$
70	4.3	$1.9 \cdot 10^{-3}$	2.3	$0.8 \cdot 10^{-1}$	5.4	$1.9 \cdot 10^{-3}$
80	1.4	$0.6 \cdot 10^{-3}$	7.4	$2.5 \cdot 10^{-2}$	1.7	$0.6 \cdot 10^{-3}$
90	4.3	$1.9 \cdot 10^{-4}$	2.3	$0.8 \cdot 10^{-2}$	5.4	$1.9 \cdot 10^{-4}$
100	1.4	$0.6 \cdot 10^{-4}$	7.4	$2.5 \cdot 10^{-3}$	1.7	$0.6 \cdot 10^{-4}$

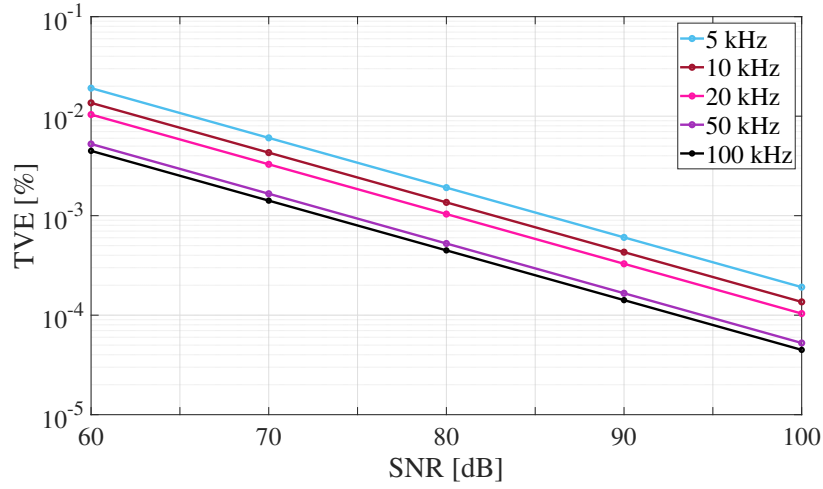


Figure 2.8: Maximum TVE for TFM and TFM_{WRLR} algorithms with different SNRs and sampling frequencies.

These considerations are fully confirmed by the simulation results. Table 2.5 reports the maximum (Max) and RMS values of TVE, FE and RFE obtained with TFM and TFM_{WRLR} methods (values are identical), which exhibit a linear increase with noise variance. Furthermore, Fig. 2.8 shows the maximum TVE achieved by both TFM_{WRLR} and TFM as a function of the SNR for different sampling frequencies. As expected, the TVE has a linear trend in the log-log plot, being it proportional to the noise standard deviation. Moreover, the graph shows that, as the sampling rate increases, the impact of SNR on TVE reduces proportionally to $\sqrt{T_s}$.

2.6.3 Amplitude and phase step tests

As previously discussed, the main advantage of the proposed methods consist in their capability of minimizing the estimation errors resulting from filter dynam-

ics triggered by abrupt transients. In order to prove that, a careful evaluation of the step change tests of the PMU Std is carried out both without and with additive noise. The first case enables verifying the validity of the mathematical treatise under ideal conditions. The second case assesses the robustness of the proposed method in the presence of wideband disturbances, other than evaluating possible degradation of the equivalent noise bandwidth introduced by the merging process. Tests have been carried out with both positive and negative steps; since the obtained results and trends are virtually identical, the analysis reported here will be limited to positive steps.

In this context, Figs. 2.9 and 2.10 refer to a positive 10 % amplitude step, occurring at $t = 1$ s, applied to TFM and TFM_{LR} algorithms. For the sake of conciseness, the figures present only the time evolution of the TVE in the noiseless and noisy case, respectively. Nonetheless, similar considerations apply also to FE and RFE.

More precisely, the upper subplot compares the TVE provided by the traditional TFM and the proposed TFM_{LR}, while showing the 1 % threshold used to define the response time. The lower subplot indicates the type of expansion selected by the TFM_{LR} algorithm. This could be either “Full”, i.e., on the entire window, or “Right” or “Left”, i.e., on a single side.

In both noiseless and noisy case, the detection criteria work perfectly. As long as the considered window does not contain the step, the TFM expansion covers the full window. Conversely, when the step occurrence enters the right side of the window, the TFM expansion is computed only on the left side, and viceversa. In this way, it is possible to keep the TVE way below the steady-state limit. As a result, the traditional TFM requires 30.1 ms to come back within the 1 % threshold, whereas the proposed TFM_{LR} has an ideal zero response time.

In the noisy case, the TVE may reach 0.012 %. If compared to the nominal conditions test, where maximum TVE is 0.003 %, such a value represents a performance degradation by about four times (yet, it remains negligible for the step change test). This is due to the fact that the expansion is computed on a shortened window, having worse equivalent noise bandwidth.

Fig. 2.11 and 2.12 are relative to a 10° positive phase step, occurring at $t = 1$ s. Differently from the amplitude case, the additive noise causes the detection criteria to fail in a few samples. This happens as long as the step instant is very close to the reporting instant, where the proposed method erroneously applies the TFM expansion on the entire window. As a result, in these time instants exactly the same performance as the traditional TFM method is achieved. Nevertheless, this has a negligible impact in terms of response time: 0.5 ms of TFM_{LR} versus 66.4 ms of TFM. This small artifact has been made visible by selecting an abnormally high reporting rate. With a lower reporting rate (in

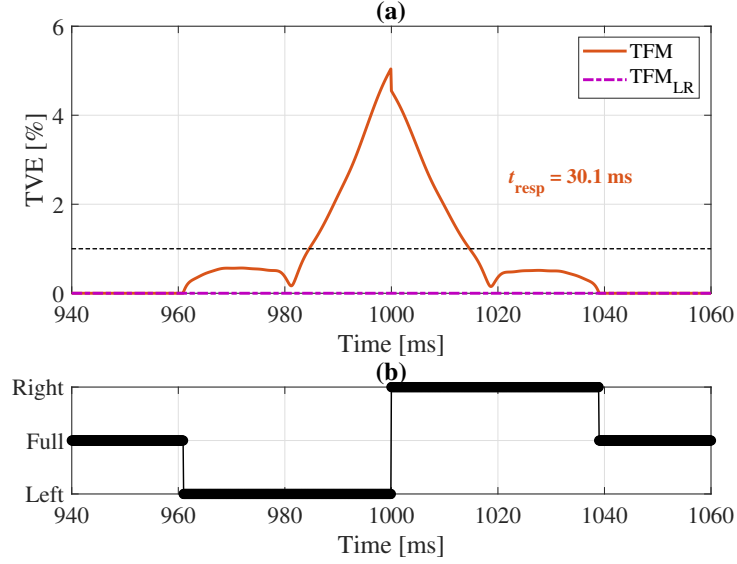


Figure 2.9: Amplitude step test in noiseless conditions. TVE as function of time as produced by TFM and TFM_{LR} in red and purple, respectively (a). Flag for left, right or full TFM expansion (b).

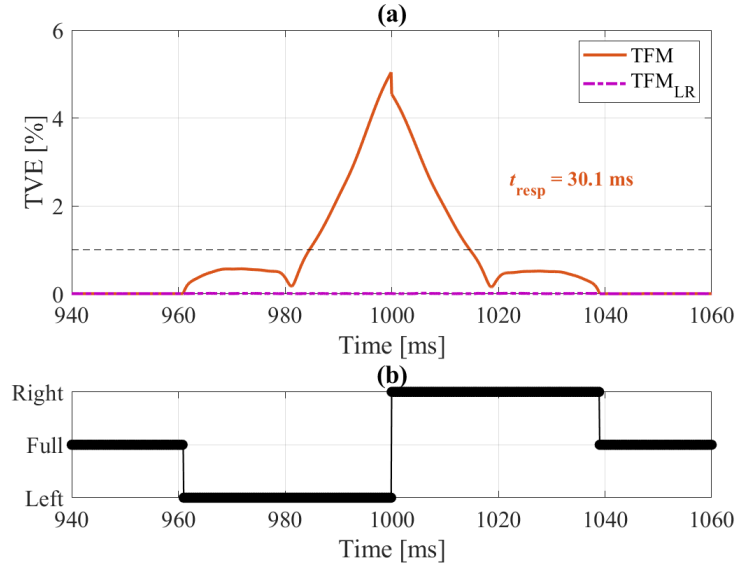


Figure 2.10: Amplitude step test with $\text{SNR} = 80 \text{ dB}$. TVE as function of time as produced by TFM and TFM_{LR} in red and purple, respectively (a). Flag for left, right or full TFM expansion (b).

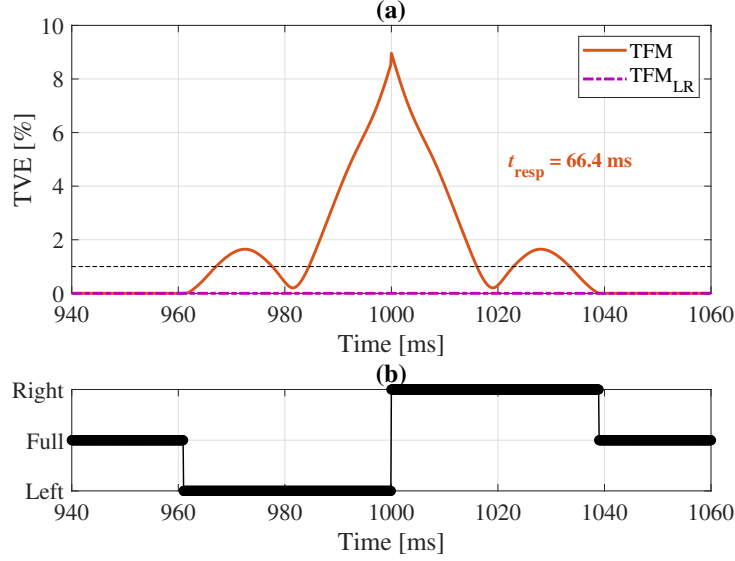


Figure 2.11: Phase step test in noiseless conditions. TVE as function of time as produced by TFM and TFM_{LR} in red and purple, respectively (a). Flag for left, right or full TFM expansion (b).

the order of tens fps, as suggested by the PMU Std), the probability of stepping into this situation is extremely low.

As a final comment regarding the proposed TFM_{LR}, it is interesting to observe that in all the step change tests the algorithm has no overshoot or undershoot, and it achieves negligible delay time. These remarkable results confirm that the proposed TFM_{LR} allows to contain the impact of transient events on PMU P-class estimation accuracy and responsiveness.

Moving to the step tests of the TFM_{WRLR}, a 10% amplitude step under noiseless conditions is again considered. In this case, the TVE of the conventional TFM implementation reaches 5% with the response time exceeding 42.51 ms. Conversely, TVE values are negligible if the TFM_{WRLR} approach is adopted, resulting in zero response time since the 1% steady-state limit is never reached. The parameter λ initially starts from small values, thus estimation is performed on the full window such as in the classic TFM approach. However, its value rapidly decreases to -1 as long as the step enters in the right half window (recombination discards automatically this part), jumps to 1 when the step moves to the left part (measurements fully rely on the right half window) then it decreases again to zero when the step is located in the rightmost samples of the window. Similar conclusions can be drawn when looking at the results in terms of FE and RFE, with the TFM algorithm reaching peaks that exceed 20 mHz and 0.55 Hz/s, respectively, while the corresponding response times are

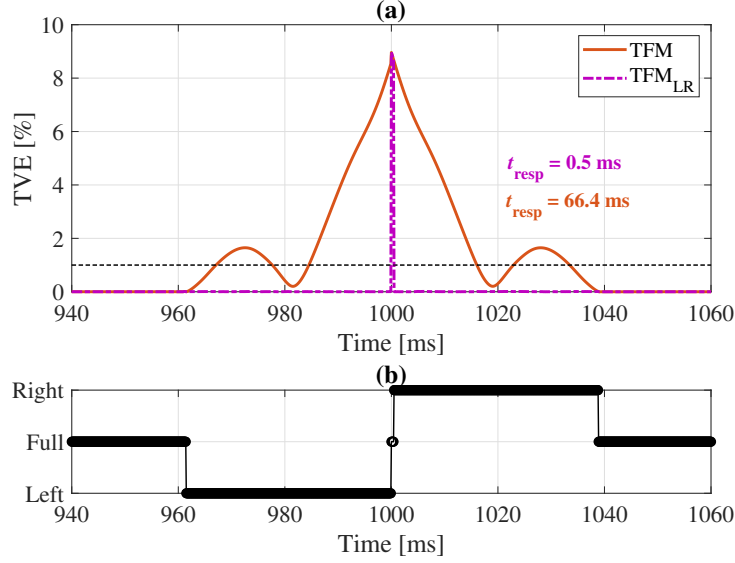


Figure 2.12: Phase step test with SNR = 80 dB. TVE as function of time as produced by TFM and TFM_{LR} in red and purple, respectively (a). Flag for left, right or full TFM expansion (b).

about 95 ms and 138 ms. Conversely, the TFM_{WRLR} technique ensures negligible FE and RFE, therefore zero response times also for frequency and ROCOF measurements.

The TFM_{WRLR} method reaches exemplary performance also in the presence of the 10° phase step, with negligible TVE, FE and RFE values, corresponding to zero response times. In this respect, the upper subplot of Fig. 2.13 reports the time evolution of the TVE for both the considered algorithms and also for the estimates obtained using only the left and right sub-windows (TFM_L and TFM_R), respectively, with the step occurring in $t = 1 \text{ s}$. The maximum TVE is well above 8% for the TFM technique, with a response time of nearly 50 ms; it is worth noting that the plot shows local maxima that are not negligible with respect to the steady-state limit, located about 45 ms before and after the step instant. The TVEs of the TFM_R and TFM_L allows a better understanding of the benefits obtained recombining by means of appropriate weights the results of the left and right TF expansions. The perfectly mirrored TFM_R and TFM_L curves exhibit almost zero TVE in the time intervals after and before the step instant, respectively, a maximum TVE of about 18% and some smaller local maxima in the mirrored intervals. Thus, they have almost ideal behavior when their contribution to the merged estimate is dominant and vice-versa.

As in the case of TVE, the classic TFM also results in extremely high peaks in term of FE and RFE values, above 600 mHz and 10 Hz/s, with response times longer than 170 ms.

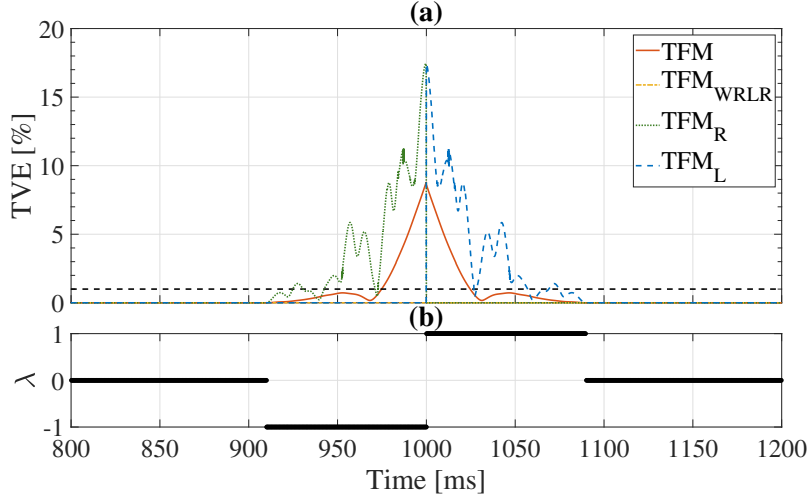


Figure 2.13: Phase step test in noiseless conditions. Time evolution of TVE for TFM (solid red), TFM_{WRLR} (dash-dotted yellow), TFM_R (dash-dotted green) and TFM_L (dash-dotted blue) algorithms (a). Time evolution of λ (b).

In addition, the lower subplot of Fig. 2.13 shows the time evolution of λ , which blends the results of the left and right expansions in the TFM_{WRLR} method. The behavior is similar to that described for the amplitude step, with λ close to 0 as the step is not within the processed time interval. When the step occurrence enters the right side of the window, λ is nearly -1, meaning that the results of the TFM_{WRLR} are practically the same of the left expansion. Conversely, when the occurrence of the step enters the left side of the window, λ is close to 1, thus the results of TFM_{WRLR} are practically the same as those of the right expansion. In this way, it is possible to keep the TVE way below the steady-state limit.

As already done for TFM_{LR}, the previously described step tests have been repeated with 80 dB SNR. In this respect, the TFM method always relies on the full vector of the samples: therefore, thanks to the averaging effect, estimates shall be rather immune with respect to noise. On the contrary, the TFM_{WRLR} may discard N samples out of $2N + 1$: this unavoidably increases noise infiltration. In turn, it may jeopardize frequency and ROCOF estimates, notably the most affected by noise since the estimation algorithm is asked to compute derivatives. Results confirm that, as expected, the performance achieved with the TFM method are barely affected by noise, both in terms of maximum errors and in response times. However, also the behavior of the TFM_{WRLR} method in the presence of step variations has proven to be robust with respect to wide-band noise. In all the cases, TVE, FE and RFE values remain very small when

compared to the corresponding steady-state limits prescribed by the PMU Std. Therefore, zero response time is attained also under these more demanding (and realistic) operating conditions.

As an example, Fig. 2.14a shows the trend of the FEs of both the methods, with an amplitude step occurring in $t = 1$ s. It is possible to see that, as the FE of the TFM method starts to rise, there is a non-negligible increase of that achieved by the TFM_{WRLR} technique. In this case, the relative weights of the left and right sub-windows are becoming very different, hence the dynamically-adjusted FIR filters have worse equivalent noise bandwidth under these conditions. Nevertheless, FE remains below 0.7 mHz. With respect to the noiseless case in Fig. 2.13, it is worth noticing how the lambda evolution (see Fig. 2.14b) presents, when the step is not inside the observation window, a noisy trend (spanning from 0.08 to -0.07). In the presence of step changes, λ shows a slightly less abrupt transition when compared to the noiseless case. The reason is that noise produces virtually identical additive contributions to $\|\mathbf{r}_{w,R}\|^2$ and $\|\mathbf{r}_{w,L}\|^2$ (proportional to its variance) that reduces $|\lambda|$. Nevertheless, the proposed method is still capable of precisely detecting the step occurrence (within few samples) and selecting the most suitable portion of the signal window.

Considering the phase step at 80 dB SNR, Fig. 2.15a reports the time evolution of the RFEs obtained by the considered algorithms. The same behavior previously observed in Fig. 2.14a here becomes more noticeable. ROCOF computation is based on a second-order differentiation, thus being even more sensitive to wideband noise. The increase of RFE that occurs when the TFM_{WRLR} method starts trusting only one of the sub-windows is evident, but its peak value remains an order of magnitude below the steady-state limit, whereas TFM strongly suffers step infiltration. When compared to Fig. 2.14b, in the presence of step changes, λ (see Fig. 2.15b) exhibits a smoother trend, but this does not affect the effectiveness of TFM_{WRLR}.

In addition to the tests performed for TFM_{LR}, additional tests were performed to characterize even better the TFM_{WRLR}, which, being proposed for M-class, requires even more accurate measurements. For this reason, the robustness of the proposed TFM_{WRLR} has been assessed through phase and amplitude steps with higher noise level. In this respect, 72 dB SNR with $f_s = 50$ kHz has been considered, thus mimicking a 12 bit acquisition stage like, e.g., in [87]. Results show that the response times to amplitude and phase steps are virtually identical to those obtained in noiseless conditions; in particular, the TFM_{WRLR} method still features zero response time. As an example, the higher subplot of Fig. 2.16, reports the RFE achieved by TFM and TFM_{WRLR} during an amplitude step. The lower subplot of Fig. 2.16 shows the evolution of λ , which is less sharp with respect to the previous amplitude step case of Fig. 2.14, but this

CHAPTER 2. ALGORITHMS FOR SYNCHROPHASOR ESTIMATION IN PRESENCE OF HIGH DYNAMICS AND TRANSIENTS

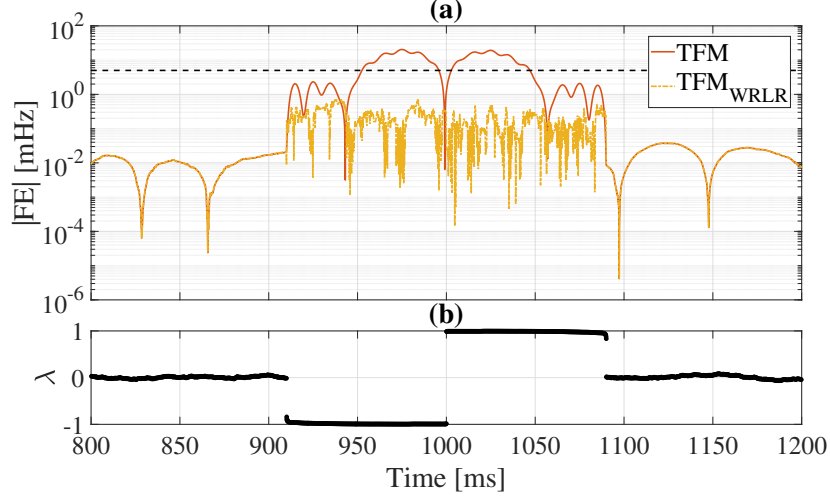


Figure 2.14: Amplitude step test with 80 dB SNR. Time evolution of FE (semilog scale) for TFM (solid red) and TFM_{WRLR} (dash-dotted yellow) algorithms (a). Time evolution of λ (b).

does not impact on the overall performance.

As a further addition of TFM_{WRLR} characterization, the step tests with 80 dB SNR have been repeated when the transition between the two steady states has finite slope. In particular, a fast but not nil transient duration is considered, where the quantity of interest (magnitude or phase angle) linearly varies between pre- and post-transient values in Δt milliseconds. Table 2.6 reports the response time T_R obtained in repeated tests with different Δt values. TVE response time is almost constant both for TFM and TFM_{WRLR}, corresponding to the value already found with instantaneous step ($\Delta t = 0$ ms in the table). Indeed, TFM_{WRLR} response time slightly increases since few measurements may exceed the steady-state threshold limit ($\Delta t = 4$ ms). This is due to the reduced leverage of balancing in the middle of the window, coupled with the decreased sensitivity of the weights in the presence of slower transients. Anyway, there is also the compensating effect that a TFM model better represents a less steep transition and this justifies the zero response time for $\Delta t = 8$ ms. As for FE and RFE, the behavior of the two methods is opposite: while TFM reduces the response time when Δt increases (the TFM model is closer to the signal dynamics), TFM_{WRLR} starts showing a less prompt reaction. Once again, the steady-state error thresholds can be exceeded near the transition, and this effect is magnified by a longer duration. Obviously, the behaviors of TFM and TFM_{WRLR} approaches tend to converge for a long transient duration since, in the presence of slow variations, their measurements are the same. However, as indicated by Table 2.6 results, with $\Delta t = 8$ ms the response times of TFM_{WRLR}

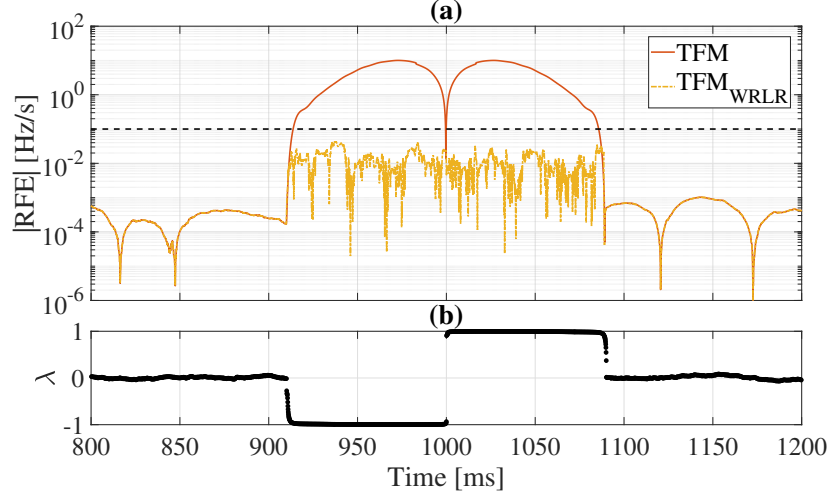


Figure 2.15: Phase step test with 80 dB SNR. Time evolution of RFE (semilog scale) for TFM (solid red) and TFM_{WRLR} (dash-dotted yellow) algorithms (a). Time evolution of λ (b).

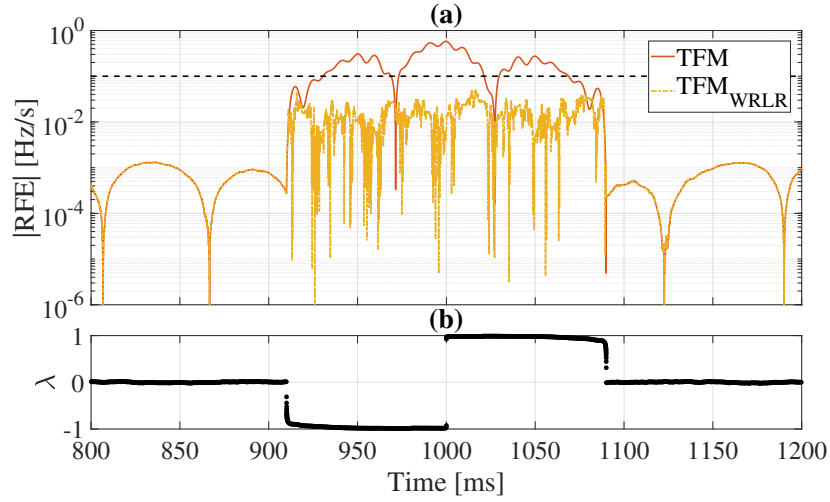


Figure 2.16: Amplitude step test with 72 dB SNR, $f_s = 50$ kHz. Time evolution of RFE (semilog scale) for TFM (solid red) and TFM_{WRLR} (dash-dotted yellow) algorithms (a). Time evolution of λ (b).

CHAPTER 2. ALGORITHMS FOR SYNCHROPHASOR ESTIMATION IN
PRESENCE OF HIGH DYNAMICS AND TRANSIENTS

are still more than seven times smaller than those achieved by TFM for both FE and RFE.

As a final comment, it is worth mentioning that response time value is strongly linked to the chosen threshold level, but, as it is patent from Figs. 2.13-2.16, the overall quality of the measurements provided by TFM_{WRLR} , when compared with a filtering algorithm like TFM, can be even better than what emerges from the mere indexes chosen by the PMU Std and this corresponds to the aim of the proposal, i.e. guaranteeing many more valid and accurate measurements notwithstanding transient conditions.

Table 2.6: Response Time of TVE, FE and RFE for TFM and TFM_{WRLR} in Linear Amplitude Transitions of Different Duration

Step Duration Δt [ms]	$T_{R,\text{TVE}}$ [ms]		$T_{R,\text{FE}}$ [ms]		$T_{R,\text{RFE}}$ [ms]	
	TFM	TFM_{WRLR}	TFM	TFM_{WRLR}	TFM	TFM_{WRLR}
0	42.5	0	94.6	0	138.4	0
4	42.0	0.2	63.7	3.3	101.4	3.6
8	41.9	0	51.8	7.1	53.2	7.5

In the next Section, an analysis of the computational burden for the proposed TFM_{WRLR} and of the response time compared with other state-of-art PMU algorithms will be reported. Similar considerations can also be applied to the TFM_{LR} , but are not presented in order to keep the discussion detailed but not too heavy.

2.6.4 Analysis of computational burden

When dealing with conventional TFFs, the computational load corresponds mainly (excluding frequency and ROCOF computation) to the application of the filter bank needed to estimate the synchrophasor derivatives. The load is thus $\mathcal{O}(M)$, since, considering the filters as pre-computed, it is due to the application of 3 complex filters to M samples ($3M$ pairs of multiplications because samples are real-valued) to find the synchrophasor and its first two derivatives. The TFM model defines the TFFs including different components in the model (4 in the above tests) and thus the computational cost is the same as the classic TFF.

In the TFM_{WRLR} method instead, the window is split into two $N + 1 = \frac{M+1}{2}$ sample segments, with one sample overlap. Thus, applying the left and right filters has almost the same cost as the usual TFFs: $3 \cdot \frac{M+1}{2} \cdot 2 = 3M + 3$ pairs of multiplications, i.e. only 3 additional pairs, regardless the value

of M . The computation cost also includes that due to (2.43), which is the estimates blending equation, namely weights computation and recombination. The weights depend on λ and thus the cost is $\mathcal{O}(N)$, because it is related to the energy computation of the left and right residuals (see (2.38) and (2.44)). Considering the main steps, there is a matrix computation, a matrix inversion, a vector computation and a final matrix-vector multiplication.

The matrix inversion is applied to a $N_p \times N_p$ matrix (where $N_p = \sum_{q=0}^Q (2K_q + 2)$ is the number of parameters in the TFM model, i.e. in $\bar{\mathbf{p}}$). In the considered implementation, the matrix is 20×20 and therefore the inversion computational complexity is $\mathcal{O}(1)$ since it does not depend on M but only on the number Q of components in the model and their degrees K_q . This is an advantage, since the matrix inversion, which could otherwise become burdensome, has indeed a limited impact if the TFM does not include many components. The matrix to be inverted is the weighted sum of the 3 pre-computed matrices $\bar{\mathbf{G}}_{w,S}$ with $S \in \{L, C, R\}$ and thus cost is $\mathcal{O}(1)$. It is indeed interesting to notice that only some rows of the inverse matrix need to be obtained, since just three estimates are necessary to compute phasor, frequency, and ROCOF of the fundamental component. Finally, vector $\alpha_L^2 \bar{\mathbf{G}}_{w,L} \hat{\mathbf{p}}_L + (1 - \alpha_L^2 - \alpha_R^2) \mathbf{b}_{w,C} x_{w,C} + \alpha_R^2 \bar{\mathbf{G}}_{w,R} \hat{\mathbf{p}}_R$ in (2.43) requires again $\mathcal{O}(1)$, since it is the sum of three matrix-vector multiplications depending only on N_p and not on M . Without considering any optimization, the number of complex multiplications in this step is $2N_p(2N_p + 4) + 1$. Then $3 \cdot N_p$ complex multiplications are required for the final matrix-vector multiplication (limiting the calculation of $\hat{\mathbf{p}}$ to the parameters of interest).

It is important to highlight that no optimization relying on the symmetry of the matrices and the presence of conjugate numbers is considered for the sake of brevity and to focus only on the essential concepts. However, it is worth mentioning that, by trading memory with computation, it is also possible to speed up the processing by pre-computing and storing the intermediate matrices corresponding to discretized values of λ , thus avoiding any matrix inversion. From the above considerations it is clear that the proposed method has a computational load comparable with that of TFFs and TFM.

Furthermore, to prove the feasibility of the method, the execution times of the TFM and the TFM_{WRLR} algorithms have been measured in a computer with Intel-Core i7-11370H at 3.30 GHz and 16 GB RAM. Considering 50000 measurements, average processing times of 67.7 and 119.4 μs have been obtained, respectively, hence guaranteeing every standard reporting rate. It is important to highlight that also in this case the symmetry of the matrices and the presence of conjugate numbers have not been exploited to reduce the computational complexity. For this reason, the previous results are only illustrative since they can be significantly improved by simplifying the algorithms operations.

2.6.5 Response time comparison of state-of-art PMU algorithms

In order to better quantify the advantage of the proposed approach, a comparison between the response times achieved by several M-class PMU algorithms published in the scientific literature, and by the reference M-class method proposed by [118] (denoted as **IEEE 60255-118-1:2018 M** method in the following description) was carried out. All of them are configured for a RR of 50 or 60 fps and are also chosen as representative of different classes of algorithms; some reach simultaneous P- and M-class compliance, thanks to different expedients aimed at reducing latency. They are briefly introduced in the following list⁴:

- **IREQ FIR-M Non-Causal**: In [119], non-causal and causal versions of a bandpass FIR filter (a Taylor window is used in the tests) with center frequency tuning are proposed. The non-causal algorithm (obtained through timestamp compensation) is designed to be M-class compliant for synchrophasor estimation.
- **M-MW-FIR**: The algorithm presented in [120] is based on the interpolation of the components extracted with two bandpass FIR filters designed starting from the Morlet wavelet and centered to rated frequency.
- **PMU Algorithm P+M**: The algorithm in [105], as already mentioned in Section 2.3, includes two TFFs (with different degrees and lengths) and a detector to switch between them in order to have simultaneous P- and M-class compliance (except for RFE out-of-band limit of [121], however removed in 2014 amendment and in PMU Std [118]).
- **i-IpDFT P+M**: This method, discussed in [84] is based on an enhanced 3-point IpDFT that compensates the long range leakage due to the negative frequency image. Location and removal of possible out-of-band components allows complying with both P- and M-class requirements.
- **HT-IpDFT P+M**: This method, proposed in [86], similarly to that in [84], is based on a 3-point IpDFT and iterative removal of possible components below 150 Hz, but it processes the signal after HT application, using an event detector to switch between two HT filters outputs with different performance.
- **FiIpDFT P+M**: The technique described in [87] can be considered as an extension of the previous i-IpDFT method [84]. It performs the measurements through 3-point IpDFT, combined with an iterative estimation

⁴The adopted names are only indicative and are either those used in the referenced paper or defined for presentation purposes.

and cancellation of the interference due to out-of-band components. It achieves simultaneous P- and M-class compliance.

- **eIpD2FT M**: The algorithm presented in [88] achieves M-class compliance with a 2-cycle observation interval through preliminary detection of interferers via ESPRIT, which are suppressed by including them into a proper dynamic model of the signal.
- **Space Vector M**: This M-class method, proposed in [122] is based on the Space Vector transformation of the three-phase signal adopting a reference frame that rotates at the nominal frequency. Positive sequence synchrophasor, frequency and ROCOF estimates are obtained by filtering the magnitude and phase of the Space Vector signal.

The results of the comparison in terms of TVE response times are reported in Table 2.7. It is important to underline that there is no pretense of completeness, since a comprehensive comparison is out of the scope of this thesis. Nevertheless, the reported results can help to better frame the potentialities of the proposed idea. Indeed, Table 2.7 demonstrates the lack of algorithms that allow zero response times for TVE, which is exactly the aim of the proposed method. Similar considerations arise when looking at FE and RFE response times (not reported here), but it is worth noting that the Space Vector M [122] algorithm achieves zero response time for these quantities in the amplitude step test.

Table 2.7: Amplitude and Phase Step Response Times of the TVE for Different State-of-Art Algorithms

Algorithm	RR [fps]	$T_{R,TVE}$ [ms]	
		Amp. Step	Ph. Step
IEEE 60255-118-1:2018 M [118]	60	-*	66.0
IREQ FIR-M Non-Causal [119]	60	30.6	37.5
M-MW-FIR [120]	50	61.6	76.9
PMU Algorithm P+M [105]	50	18.0	22.0
i-IPDFT P+M [84]	50	28.0	32.0
HT-IPDFT P+M [86]	50	22.9	26.4
FiIPDFT P+M [87]	50	28.1	34.1
eIpD2FT M [88]	50	14.0	24.0
Space Vector M [122]	50	37.5	42.5
Conventional TFM M	50	42.5	50.1
Proposed TFM _{WRLR} M	50	0	0

* The PMU Std reports only the value for the phase step change test.

Chapter 3

Advanced synchronization technologies

This chapter first presents standard synchronization technologies such as the Global Navigation Satellite System (GNSS) and packet-based time Precision Time Protocol (PTP), and confronts them with the promising White Rabbit (WR) protocol, discussing especially the benefits of WR for power system applications. After that, the achievable synchronization accuracy through real-time industrial controllers using the WR technology is analyzed comparing it with Global Positioning System (GPS) results. In addition, the implications in power system applications arising from the uncertainties introduced by synchronization are discussed. The obtained results show that WR protocol could complement GPS technology in several power system applications in the future, including harmonic synchrophasor estimation and traveling wave-based fault location.

3.1 Overview of synchronization in power systems

As seen partially in previous chapters, nowadays, time-sensitive applications characterized by increasingly stringent synchronization requirements, covering ever-larger areas and an ever-increasing number of devices, are becoming essential in power systems. In this context, there is a clear need to explore new synchronization technologies that can provide better performance while ensuring improved portability and scalability in comparison to typical synchronization technologies. For this reason, the aim of this section is to give an overview of synchronization technologies, mainly demonstrating the advantages of time synchronization with the WR protocol for modern power system monitoring and

control applications.

In this context, the starting point is the findings of [123], which explored the synchronization of Phasor Measurement Units (PMUs) prototypes using various synchronization protocols, including WR. In [123], the results of PMU measurements are reported in terms of Total Vector Error (TVE) and phase-angle error, taking into account different test signals during the characterization process. These metrics were intended to show the effect of the time synchronization source and compare it to other uncertainty contributions (e.g., the estimation algorithm or the acquisition noise). This choice helps to consider the impact of time synchronization on the final application, but it can also cover the individual error contribution of the elements in the acquisition and evaluation chain. In this chapter, the focus is instead on the synchronization error of the hardware platform itself. For each considered configuration, the statistical distribution of the internal hardware time reference is characterized. Based on these information, systematic and random contributions to the synchronization uncertainty are identified.

In compliance with the PMU Std [118], as already mentioned, the target synchronization accuracy is in the order of 1 microsecond, which corresponds to a phase-angle measurement error of about 0.3 mrad at a nominal frequency of 50 Hz. To this end, PMUs typically employ a GPS receiver to tag each synchrophasor measurement with an absolute timestamp within the Coordinated Universal Time (UTC) timescale. Alternatively, Ethernet-based protocols exploit the existing communication infrastructure to distribute the synchronization among the different nodes of the network [124], [125].

From the evolution of Ethernet-based synchronization protocols was born the advanced WR protocol. The advantages of this technology are not limited to the high time accuracy that can be achieved, but they include the possibility of sending the time information over long distances [126]. This aspect can be vital in the intercommunication between remote electrical substations.

For this reason, GNSS technology, Ethernet-based synchronization time protocols, and the innovative WR technology will be discussed in detail separately in the next section.

3.1.1 Global Navigation Satellite System

GNSS is a widely used technology for geolocalization and for time synchronization, offering position accuracy within a few meters and time markings in the tens of nanoseconds range. It is based on satellites that allow sending time information from accurate atomic time sources to the ground. Hence, GNSS is, from the time perspective, an atomic time propagation systems. GNSS is a

general term for different constellations of satellites such as the american GPS, russian GLONASS, European Galileo and chinese BeiDou, which provide receivers around the world with highly accurate time and position data essential for applications requiring both precise timing and geographic positioning [127]. GNSS positioning allows operators, using just a GNSS receiver, to locate devices even while moving over large geographic areas, which is the reason why it is widely used by civilians for orientation and especially for navigation but also by commercial and military applications such as asset tracking and aviation [128].

The GNSS infrastructure is typically composed of three main segments: the space segment, the control segment and the user segment [127].

The space segment consists of a constellation of satellites typically in Medium Earth Orbit (MEO), i.e., range of $2000 \div 35786$ km altitude, which continuously transmit signals containing accurate information on the position of the satellite itself and time provided by the atomic clock. Main GNSS constellations are designed to ensure global coverage, with satellites distributed on different orbital planes.

For example, in the case of the American GPS, there are at least 24 active GPS satellites in orbit, distributed on six orbital planes to ensure global coverage and reliable 24-hour availability of the signal. The orbit is almost circular around the Earth, at an altitude of about 26600 km, with a relative speed to the ground of about 3000 km/h. The orbital period is one-half of a sidereal day (11 hours and 58 minutes local time) and the orbit belongs to 6 equidistant orbital planes, 60° separation in longitude and 55° on the equatorial plane. Each orbit has 4 satellite slots and this number, as will be discussed shortly, is the minimum to be able to obtain a reliable estimate of the time in the GPS receivers. Each satellite is equipped with two caesium and two rubidium clocks with a drift in the order of 1 ns per day, which are synchronized with the GPS ground control stations to maintain consistent timing throughout the constellation.

The control segment includes a network of ground stations that monitor, control and maintain the satellites in orbit. This includes:

- Master Control Station (MCS): the main control structure responsible for satellite tracking, health monitoring and maintenance of the system. Each GNSS constellation has a centralized control station that supervises the operations of its satellites. It receives data from fixed stations and processes them to calculate the orbital ephemeris of each satellite and to evaluate the corrections that must be periodically applied to the satellite clocks and, possibly, to the orbits. The MCS is often equipped with an atomic clock that is more accurate than those on board in the satellites.

For example, in the case of GPS, the MCS has a reference hydrogen atomic clock, 50-100 times more stable than caesium clock of the satellites, allowing clock corrections achieving a drift in the order of 100 ps in 24 hours. In the event of a satellite failure or malfunction, the MCS also takes care of temporarily or permanently excluding the satellite from the system. For redundancy, there is often also an alternative MCS that can take the place of the main one at any time.

- Control antennas: These antennas are used to communicate commands from the MCS to the satellites. They transmit periodically the MCS's instructions, including new ephemeris data and clock correction parameters, directly to the satellites to ensure they remain synchronized and in the correct orbit.
- Monitoring stations: these globally distributed stations track the satellite signals, monitor the quality of the signal and communicate all this information to the MCS. This continuous monitoring helps the MCS make accurate adjustments to satellite positioning and clock synchronization as needed.

Finally, the user segment is represented by the set of GPS receivers of the users that decode satellite signals to determine position, velocity and time. These devices are radio receivers tuned to the specific frequencies of the GNSS system, e.g., in the case of GPS two carrier frequencies are used: L1 1575.42 MHz and L2 1227.60 MHz. The receiver is therefore composed of an antenna, a microprocessor, an accurate oscillator (typically a good quality quartz oscillator), a memory and a software module to process the acquired data and provide estimates of position and time. The receiver is often able to provide a synchronization signal as output using a classic single analog voltage pulse every second, commonly called pulse per second (PPS), or more complicated signals such as Inter-Range Instrumentation Group (IRIG) timecodes [129].

Although the cost to the user who wants to use GNSS technology is quite low due to the low cost of GNSS receivers embedded in integrated circuits, the overall cost of designing and implementing a GNSS system is about tens of billions of euros, in addition to the cost of maintenance and constellation replacement every about 15 years (currently the cost of a single satellite is half a billion euros). As well as the cost, the deployment time of such systems is really high, and for these reasons the number of GNSSs is limited to few countries.

Regarding time synchronization, GNSSs enable devices such as PMUs and other critical network monitoring and control equipment to maintain time synchronization in wide areas. This is essential for obtaining accurate phase-angle measurements of synchrophasors.

The following section describes how exactly receivers compute position and time through messages received from satellites. Assuming that there are three satellites, each providing a receiver with a time-aligned signal including the satellite's position in space, it is possible to define the unknown position of the receiver as $\mathbf{P} = (x, y, z)$ and the known positions of the satellites in space as $\mathbf{U} = (U_x, U_y, U_z)$, $\mathbf{V} = (V_x, V_y, V_z)$ and $\mathbf{W} = (W_x, W_y, W_z)$. Considering the satellite U , by comparing the time instant t_{sU} at which the signal was sent by the satellite with the time it was received t_{aR} , the “flight time” of the signal Δt_U can be estimated. By then multiplying the latter by the speed of the electromagnetic wave carrying the signal (which can be well approximated by the speed of light c), the distance from $\mathbf{P}$ to satellite U can be obtained. This distance can be considered the radius R_U of a hypothetical sphere with the center on the corresponding satellite U . Repeating the same approach for all satellites, it is possible to write the equation for position triangulation as:

$$\begin{cases} (x - U_x)^2 + (y - U_y)^2 + (z - U_z)^2 = R_U^2 = (\Delta t_U \cdot c)^2 \\ (x - V_x)^2 + (y - V_y)^2 + (z - V_z)^2 = R_V^2 = (\Delta t_V \cdot c)^2 \\ (x - W_x)^2 + (y - W_y)^2 + (z - W_z)^2 = R_W^2 = (\Delta t_W \cdot c)^2 \end{cases} \quad (3.1)$$

From this system of three equations, it might appear easy to estimate the position of the receiver, but in reality there are 4 unknowns. Indeed, the receiver clock is not as accurate as the atomic clocks of the satellites, which results in an unknown time shift δt in the receiver measurements (time offset). To account for this, the measured flight time $\widehat{\Delta t_U}$ (the hat is used to highlight that this is a measured quantity) for satellite U can be rewritten as:

$$\widehat{\Delta t_U} = t_{aR}^2 - t_{sU}^1 = (t_{aR}^1 + \delta t) - t_{sU}^1 = \Delta t_U^1 + \delta t \quad (3.2)$$

where:

- t_{sU}^1 represents the absolute time in the GNSS timescale (very close to UTC) when the signal was transmitted from the satellite;
- t_{aR}^2 is the arrival time of the signal at the receiver, thus affected by the unknown time offset δt ;
- t_{aR}^1 is the arrival time of the signal at the receiver expressed in the GNSS timescale.

Similar relationships can be applied for satellites V and W . Therefore, substituting Δt_U^1 into Eq. (3.1) (and the same for the other satellites) gives a system of 4 unknowns that necessarily requires at least the information of a

fourth satellite S to be solved. Hence, the new system becomes:

$$\begin{cases} (x - U_x)^2 + (y - U_y)^2 + (z - U_z)^2 = R_U^2 = (\widehat{\Delta t}_U - \delta t)^2 \cdot c^2 \\ (x - V_x)^2 + (y - V_y)^2 + (z - V_z)^2 = R_V^2 = (\widehat{\Delta t}_V - \delta t)^2 \cdot c^2 \\ (x - W_x)^2 + (y - W_y)^2 + (z - W_z)^2 = R_W^2 = (\widehat{\Delta t}_W - \delta t)^2 \cdot c^2 \\ (x - S_x)^2 + (y - S_y)^2 + (z - S_z)^2 = R_S^2 = (\widehat{\Delta t}_S - \delta t)^2 \cdot c^2 \end{cases} \quad (3.3)$$

This system can be easily solved, thus estimating x , y , z and δt . To increase the accuracy of GNSS, it is possible, in the case of visibility of more than 4 satellites (they may also correspond to different GNSSs), to add more equations to the system and estimate the unknowns, e.g. using the least squares (LS) method, more accurately.

The accuracy achieved by GNSS depends on various factors, including the type of receiver, atmospheric conditions, and the quality of satellite signals. Under ideal conditions, GNSS can provide timing accuracy within 10-20 nanoseconds and positioning accuracy within 3-5 meters for standard receivers. More advanced systems, such as dual-frequency receivers or those using differential GNSS (DGNSS) techniques, can achieve sub-meter positioning accuracy and timing accuracy within a few nanoseconds, making them suitable for highly time-sensitive applications [130].

DGNSS techniques enhance positioning accuracy by using a network of fixed ground-based reference stations that transmit correction signals to GNSS receivers in the area. These corrections help mitigate errors caused by atmospheric delays, satellite clock inaccuracies, and other environmental factors, allowing receivers to achieve greater accuracy. DGNSS is particularly valuable in applications requiring high accuracy, such as precision agriculture, surveying, and autonomous vehicle navigation, where positioning errors must be minimized [131].

Despite the high accuracy of GNSS, the technology faces challenges, including vulnerability to environmental interference, signal blockage (e.g., in urban canyons, forested areas, inside buildings, and underground), and potential security risks such as jamming or spoofing [132]. These vulnerabilities have led to the development and integration of alternative synchronization methods, such as PTP and WR, in some applications to complement GNSS. These complementary technologies offer increased robustness in environments where GNSS reliability may be compromised. Nevertheless, GNSS remains a cornerstone of time synchronization in power systems due to its wide adoption, proven accuracy, and global availability. It is typically the primary source of synchronization in a power network monitoring system.

3.1.2 Synchronization time protocols

In the context of the IEC 61850 series of standards, which describes the communication methods and specific devices for digitized electrical substations, where measurements and control signals are digitized, device synchronization is generally performed using the IEEE 1588 protocol [35], commonly called PTP, which allows time synchronization in the order of one microsecond to be maintained within the substation. To achieve this level of synchronization, the PTP protocol is configured with a specific set of parameters, called profile, to be used in the specific context of the digital electrical substation [133]. This profile allows for a level of time accuracy in the order of microseconds considering up to 15 compatible devices [134], limiting its use to the substation.

The synchronization mechanism is similar to other synchronization protocol based on Ethernet packets, like Network Time Protocol (NTP) [135]. As PTP, NTP is also a client-server protocol used to synchronize the clocks of a distributed system on an Internet network and allows accuracy ranging from a few milliseconds to a few seconds, depending on network traffic, the performance of the device being synchronized and the operating system.

In both time protocols, the synchronization mechanism relies on a series of message exchanges between a master clock and a slave clock to achieve accurate time alignment. Specifically, in the PTP, the process consists of four key messages: “Sync”, “Follow_Up” (in two-step mode), “Delay_Request”, and “Delay_Response” [35]. This sequence is represented in Fig. 3.1 and allows the slave to calculate the offset and path delay, which are crucial for accurate synchronization.

The synchronization process begins when the master clock records a timestamp t_1 and sends it to the slave via the Sync message. This message contains the initial time reference and is used by the slave to initiate its calculations. In one-step mode, the Sync message alone provides the timestamp. In two-step mode, an additional Follow_Up message is sent after the Sync message to carry the timestamp for enhanced accuracy. This message ensures that any delays in transmitting the Sync message do not affect the synchronization process. Then, the slave clock records its own timestamp t_2 as soon as it receives the Sync message. After that, it sends a Delay_Request message back to the master, marking the time of transmission as t_3 . Upon receiving this message, the master clock records another timestamp t_4 and sends it back to the slave in a Delay_Response message. With all four timestamps, the slave can compute the master-to-slave delay T_{ms} and the slave-to-master delay T_{sm} . These delays can be expressed as follows:

$$T_{ms} = (t_2 - \delta t) - t_1 \quad (3.4)$$

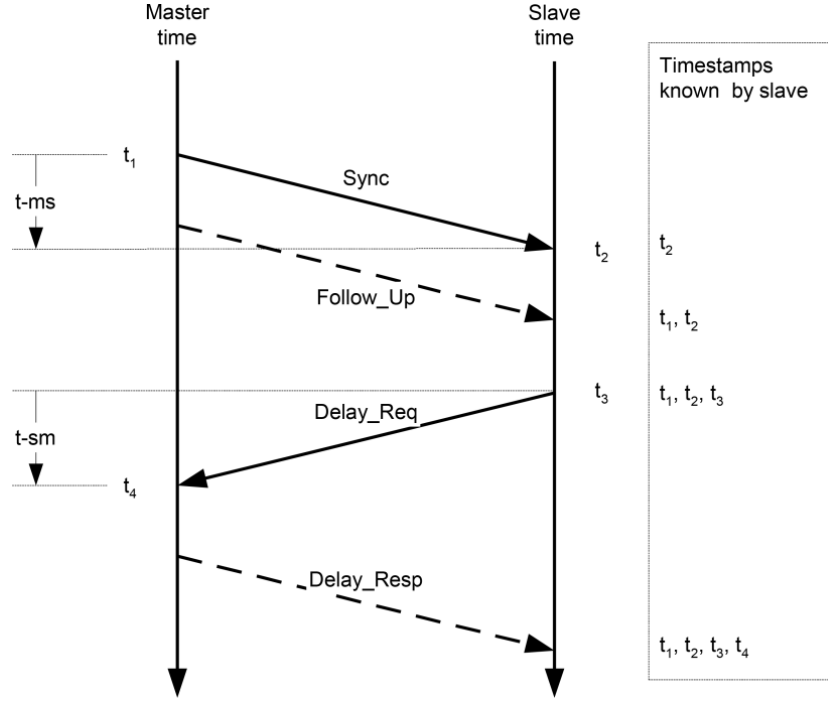


Figure 3.1: Exchange of synchronization messages between a master and a slave clock in IEEE 1588 [35].

$$T_{sm} = t_4 - (t_3 - \delta t) \quad (3.5)$$

where t_2 and t_3 are timestamps computed through the slave time clock and, for this reason, they are affected by an offset δt with respect to the master time. The offset is what needs to be estimated to synchronize the slave clock. Now, by subtracting Eq. (3.5) by Eq. (3.4) and making δt explicit, the following expression is obtained:

$$\delta t = \frac{(t_2 - t_1) - (t_4 - t_3)}{2} - \frac{T_{ms} - T_{sm}}{2} \quad (3.6)$$

Finally, by summing Eq. (3.4) with Eq. (3.5), the so-called Round Trip Time (RTT) can be calculated as:

$$T_{ms} + T_{sm} = \text{RTT} = (t_2 - t_1) + (t_4 - t_3) \quad (3.7)$$

The RTT value represents the total time taken for a message to travel from the master to the slave and back. If symmetry is present in the network path, hence the delay from the master to the slave is the same as the delay from the slave to the master, then each path delay is equal to half of the RTT, i.e.:

$$T_{ms} = T_{sm} = \frac{\text{RTT}}{2} = \frac{(t_2 - t_1) + (t_4 - t_3)}{2} \quad (3.8)$$

Under symmetry assumption the calculation of the slave offset is:

$$\delta t = \frac{(t_2 - t_1) - (t_4 - t_3)}{2} \quad (3.9)$$

just eliminating the last term of Eq. (3.6) that depends on path delays. By using these computations, the slave clock can adjust its local time to align closely with the master clock, achieving synchronization with minimal error.

In addition, PTP, as well as WR, allows to increment the time accuracy, since synchronization messages can be fully marked at the hardware level, ensuring greater determinism and reducing the contribution due to queues of operations at the software level. In addition, the PTP network requires all network devices to be able to identify synchronization messages in order to compensate for the propagation delay introduced by each network element. Standard 1588 thus defines three different classes of network devices: the ordinary clock (OC), the boundary clock (BC) and the transparent clock (TC) [35].

An OC is a PTP-enabled device that has a single PTP port and is usually configured either as a master or a slave in the PTP hierarchy. Therefore, the primary function of an OC is to generate or receive the reference time. A master OC provides the primary time source for other clocks in the network, while a slave OC synchronizes its internal clock to match the time from its master. In a PTP network, the master clock can be chosen as the grandmaster (GM), i.e., the most accurate and reliable time source, through the Best Master Clock (BMC) algorithm.

The BMC algorithm is an essential feature of PTP and is responsible for automatically choosing the most suitable clock to serve as the GM of the network. During the initialization phase, each clock in the network exchanges timing information through a series of “Announce” messages. These messages allow the BMC algorithm to compare their quality and capabilities based on factors such as clock accuracy, stability, and traceability to a primary reference time. The BMC algorithm then determines the “best” clock based on these criteria, designating it as the GM clock. If a higher quality clock is later introduced into the network or there is a significant change in network conditions, the BMC algorithm re-evaluates the hierarchy and can promote the new clock to GM status, ensuring optimal timing accuracy throughout the network. After GM selection, the actual synchronization phase begins, in which timing information is distributed to slave devices in the network. Slave OCs are typically found in end devices that require precise timing, such as measurement instruments, controllers, and other time-sensitive devices.

A BC is a multi-port device that operates at the intersection of different network segments, helping to maintain synchronization across them. Unlike an OC, a BC has multiple PTP-enabled ports, allowing it to connect to multiple

other clocks in the network. In a typical setup, one port of the BC serves as a slave port that synchronizes to the upstream master clock, while the other ports act as master ports, providing synchronized time to downstream devices. This structure allows the BC to reduce timing inaccuracies that could accumulate across long chains of devices, enhancing overall network stability and timing precision. Boundary clocks are especially useful in complex network topologies where multiple segments need to be synchronized without directly passing through the primary master clock.

A TC is a PTP-enabled device that maintains high timing accuracy by correcting the delay introduced when timing packets pass through itself. Unlike OCs and BCs, a TC does not participate in the PTP hierarchy as a master or slave but instead works as a repeater by also measuring the time it takes PTP packets to transit the device (known as residence time) and updates a correction field within the packet to account for this delay. This ensures that downstream devices can accurately account for network-induced delays, improving overall synchronization accuracy.

The PTP protocol has been gradually improved over the years, and this has led to the development of the promising WR.

3.1.3 White Rabbit protocol

The WR protocol is an advanced synchronization technology designed to address the growing demand for ultra-reliable sub-nanosecond accuracy in time-sensitive applications. It was originally developed at European Organization for Nuclear Research (CERN) to synchronize different processes in the particle accelerator complex [136], [137]. Nowadays, WR is described as High-Accuracy profile to the latest PTP version in Annex M of IEEE Standard 1588 [35].

This protocol is designed to operate over different ranges of network topologies, from simple point-to-point connections to complex mesh or tree configurations. Synchronization is spread according to a master-slave hierarchy and propagated from the top, i.e., from the WR master, to the bottom, reaching the other WR switches and finally the last nodes. WR networks can consist of thousands of devices several kilometers apart and connected via existing Ethernet networks. This adaptability makes it suitable for a wide variety of applications. Moreover, the WR Protocol is designed to scale efficiently, accommodating network growth and increasing the number of devices without sacrificing performance in terms of accuracy, thanks to the accurate estimation and correction of the delays of each node.

The WR Protocol is based on the standards:

- IEEE 802.3 for the Ethernet physical layer [65];

- Synchronous Ethernet (Synch-E) for the transference of clock signals over the Ethernet [138];
- IEEE 1588 for the PTP packet synchronization protocol, from which the WR protocol is derived [35].

Through these standards, and using a technique called Precise Phase Measurement, the WR protocol achieves remarkable performance, depending on the communication medium. The best results are reached using fiber optic connection between the nodes of the WR network, where the synchronization accuracy, defined as the deviation of each node with respect to the UTC reference, falls in the sub-nanosecond range with picoseconds precision [139]. Instead, through conventional Ethernet over copper, it is possible to achieve an accuracy of about 30 ns, which is almost two orders of magnitude worse than in the former case [140]. This substantial improvement arises from the many advantages of optical fiber including better immunity to electromagnetic interference and higher bandwidth, and from that of Sync-E, which has a clock accuracy of ± 4.6 ppm and thus represents a great improvement over the asynchronous Ethernet limit with a free-running clock having ± 100 ppm accuracy [141].

This innovative technology provides a suitable and efficient substitute or enhancement with respect to GPS, especially in scenarios where (i) sky access is limited, such as in urban environments, (ii) there is pre-existing telecommunication infrastructure, and (iii) the typical distance between the nodes is below 10 km [126]. Moreover, the WR synchronization is more stable and less geographically dependent than GPS.

Regarding the power systems, in addition to the remarkable accuracy, the WR represents a significant advantage in the substation as it allows the transition from synchronization at the local level (as in the classic GPS), to a single source of wide-area synchronization that can be used in several substations and re-transmitted between them. To highlight this important concept, Fig. 3.2 depicts two typical synchronization architectures based on GNSS on the left and WR technology on the right.

Time synchronization accuracy requirements in power systems depend on the specific application and IEC Standard 61850-5:2013 [142] defines six classes (see Table 3.1), ranging from Class A for phasor and distributed measurement applications (accuracy $\pm 1 \mu\text{s}$) to Class F for no time-critical monitoring applications ($> 1 \text{ s}$).

Even Class A synchronization accuracy may not be sufficient for advanced time-sensitive applications, such as high-order harmonic synchrophasor measurements.

In the context of digital substations based on the IEC 61850 standards,

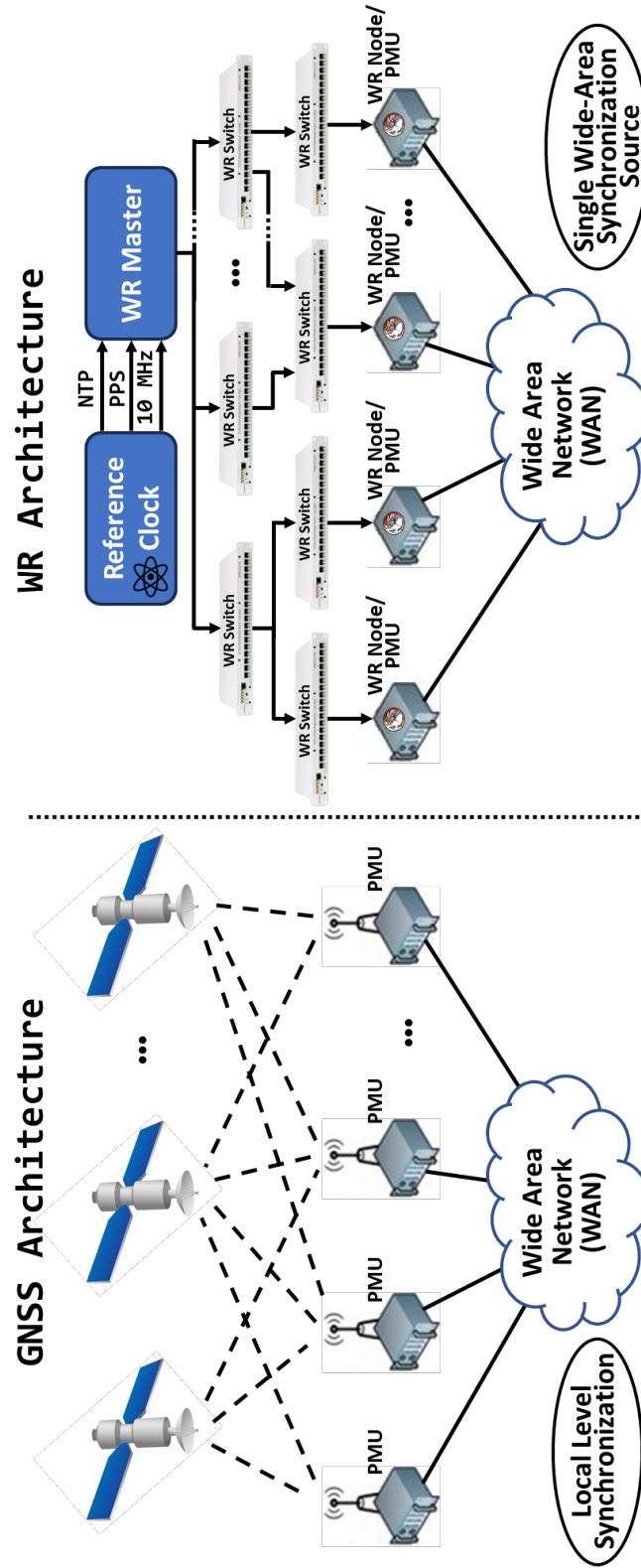


Figure 3.2: Typical architectures based on GNSS on the left and on a single wide-area synchronization source with WR technology on the right.

Table 3.1: Time Synchronization Accuracy Classes for Power Systems Applications in IEC Standard 61850-5:2013

Class	Synchronization Accuracy	Applications
A	1 μ s	Phasor and distributed measurements
B	100 μ s	Automated fault recording
C	1 ms	Time tagging of fast/transient events
D	10 ms	Power quality monitoring
E	100 ms	Slow event monitoring and logging
F	< 1s	Other no time-critical monitoring applications

the IEC Rootable Generic Object Oriented Substation Event (R-GOOSE) and Rootable Sampled Value (R-SV) [143] protocols allow the communication of measurement and control signals between substations thanks to their ability to be encapsulated at higher layers; e.g., data streams encapsulated in Transport Control Protocol (TCP) or User Datagram Protocol (UDP) data packets. Several examples explore the possibility of sharing measurement and control signals between different substations using R-GOOSE and R-SV signals [144]. In this context, time synchronization can also be shared between substations, and the WR protocol is a candidate for wide-area applications, because it allows synchronization of points that are far apart without losing temporal accuracy. The advantage of considering a single synchronization source is, for instance, to overcome offset issues occurring between devices installed in different substations, or to prevent time synchronization degradation in a single substation from bringing difficulties in correlating data from various substations. In line with these considerations, the installation of fiber networks for non-strictly research purposes is rapidly growing (e.g., SWITCH [145] and LIFT [146] projects in Switzerland and Italy). In fact, using fiber optics, it is possible to transport the WR protocol tens of kilometers without the need to retransmit the signal. As happened with Generic Object Oriented Substation Event (GOOSE) and Sampled Value (SV) messages that evolved to be routable outside the substation [143], the synchronization signal could also follow the same evolution in the future.

3.1.4 Examples of time-sensitive power system applications

There are different power system applications that can benefit from a better synchronization. In this thesis, two relevant applications are considered: the extension of synchrophasor measurements to harmonic tones and the identifica-

tion of traveling wave arrival times for fault location.

In the first case, the aim is to design instruments able to measure magnitude and phase-angle also for the higher-order harmonics, generally up to the 50th as suggested by the IEC Standard 61000-4-7 [147]. Refined algorithms have shown promising results [148], [149], but mostly based on performance characterization in simulated environments that do not account for synchronization errors, rather than implementing prototype devices. In such applications, the time synchronization accuracy is crucial for estimating phase angles of high-order harmonics. In this context, it is essential to evaluate advanced synchronization methods that can provide accuracy at the nanosecond level, which is the target of the WR protocol.

In the second case, fault location typically relies on the comparison of the arrival time of two traveling waves. A more accurate synchronization would reduce the measurement error and produce a more accurate and reliable location.

Given the relevance of synchronization in this context, the contribution of the implemented synchronization to measurement error for these two applications will be analyzed in detail in Section 3.4.

3.2 Implementation of WR architectures

This section initially discusses two architectures based on WR and modular systems that have been proposed in [150] and the test setup used for their characterization, then it presents the obtained results.

Fig. 3.3 presents the two architectures and the testing scenarios. Each of them is based on National Instruments (NI) compactRIO (cRIO) system, which consists of an embedded industrial controller equipped with a real-time (RT) processor running NI Linux RT operating system. These controllers are equipped or can be equipped with a chassis featuring a user-programmable Field Programmable Gate Array (FPGA), which allows deterministic and high-speed communications with the added I/O modules.

The first architecture (Fig. 3.3(a)) was implemented in the Electrical and Electronic Measurements Laboratory of the University of Cagliari. The first element of the architecture, the Meinberg Lantime M1000 [151], is a GPS receiver that provides a PPS accuracy better than ± 50 ns. This device allows synchronization to be distributed to external systems by generation of PPS signal, a 10 MHz square wave and also through synchronization packets using, for example, NTP. These three signals were used to synchronize the WR Switch from SAFRAN [152], which allows synchronization to be distributed to over a thousand nodes with sub-ns time accuracy via optical fibers using the WR protocol. For this architecture, in particular, the NI cRIO-9024 controller [153] was used,

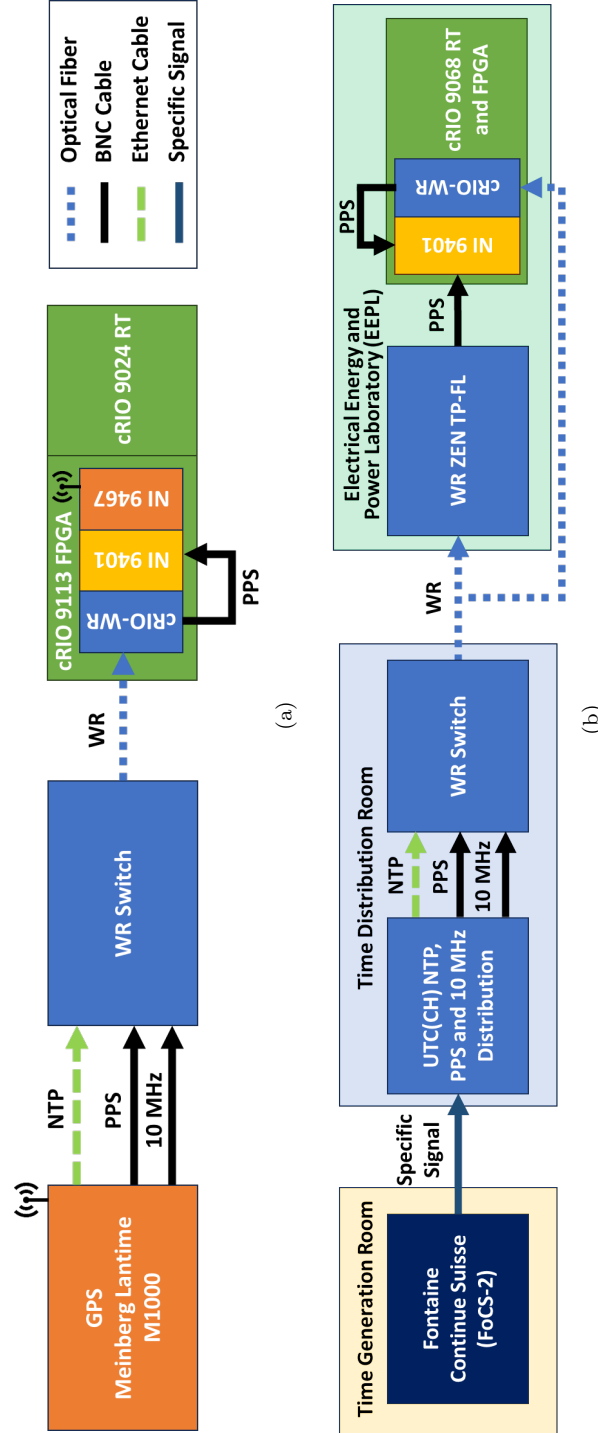


Figure 3.3: Schematics of the experimental test architectures implemented in the Cagliari laboratory (a) and in the METAS laboratories (b).

which is equipped with a RT processor having 800 MHz frequency and 512 MB of DRAM memory. This controller was provided with the NI cRIO-9113 chassis [154] that featured a Virtex-5 LX50 FPGA characterized by a maximum time-base accuracy of 100 ppm and 250 ps maximum jitter for 40 MHz frequency. Three boards were placed inside the chassis:

- NI 9467 is a GPS synchronization module that provides PPS accuracy of ± 100 ns. This module is used in the tests to have a purely GPS-based architecture for comparison purposes;
- NI 9401 is a digital I/O interface with maximum propagation delay of 100 ns;
- cRIO-WR is a module equipped with a WR core implemented on the Xilinx Spartan-6 FPGA [155]. It provides a Small Form-factor Pluggable (SFP) cage for fiber-optic transceiver used for sending and receiving the WR protocol, a USB connector for user configuration and an HDSUB-15 port with programmable I/O.

The last module was configured in slave mode and is point-to-point connected via a fiber-optic link to the WR Switch, operating in grandmaster mode [152]. In addition, the cRIO-WR module was configured to generate in the I/O port a PPS signal directly connected to the NI 9401 via BNC cable.

Instead, the second architecture (Fig. 3.3(b)) was implemented in the Swiss Federal Institute of Metrology (METAS) laboratories in Bern-Wabern and it employs a system capable of guaranteeing traceability to UTC(CH). The time reference is generated within the Time Generation Room via the Fontaine Continue Suisse (FoCS-2), which represents the primary frequency standard, ranked among the world's most accurate caesium atomic clocks. It deviates by only one second from the exact time over 30 million years, so it achieves time accuracy on the order of femtoseconds [156]. The reference provided by FoCS-2 is distributed through the Time Distribution Room where, as seen for the Meinberg GPS receiver, a PPS signal, a 10-MHz clock, and NTP protocol are generated in order to synchronize the WR Switch. Lastly, the WR Switch is set as the grandmaster of the WR network, so it is responsible for transmitting the synchronization via WR protocol to the next nodes with sub-nanosecond accuracy. The WR network is composed of two WR slaves: the cRIO-WR module already presented in the architecture implemented in Cagliari and a higher performance device, the WR ZEN TP-FL [157], calibrated by METAS at 7 ps by UTC(CH). Both devices allow generating a PPS that is sent as input to the NI 9401 digital board as in the first architecture. In this case, the NI cRIO-9068 [158] equipped with ARM Cortex-A9 dual-core processor with a maximum frequency of 667



Figure 3.4: Block diagram of the software implemented to synchronize NI cRIO devices with PPS.

MHz and 1 GB RAM was used as the RT controller. This modular system is also already provided with a chassis including a Xilinx Zynq 7020 FPGA with a frequency of 40 MHz and a RT clock with a guaranteed accuracy of ± 5 ppm. Within the chassis, the NI 9401 digital board and the cRIO-WR module were included in the METAS architecture.

The software component implemented in the NI cRIO-9024 and 9068, of which a block diagram is shown in Fig. 3.4 will be described below. This software has been developed in LabVIEW environment for both controllers to disciplines the internal clock of the FPGA with the two proposed synchronization methods. Hence, the FPGA internal clock is synchronized via a PPS signal using the FPGA Timekeeper library, continuously acquiring the PPS via digital input from the NI 9401 or via the NI 9467 GPS receiver. Whenever a rising edge is detected, the Timekeeper timestamp (i.e., the timestamp provided by the library) is saved and the difference from the time reference is made by calculating the offset of the two clocks. Finally, the FPGA clock is updated and the offset value is sent to the RT controller, which is responsible for the data file storage.

3.3 Characterization of synchronization errors

This section describes the results of four tests performed using the two architectures and the previously presented software. Considering the architecture in Fig. 3.3(a), two tests were carried out: in the first, which will be referred to in brief as “GPS” test case and represents the typical synchronization used nowadays, the NI 9467 GPS board is used for synchronization; in the second, which will be called “GPS-WR” test case, the PPS coming from the cRIO-WR module is exploited (note that the WR Switch is synchronized via the Meinberg GPS receiver). The other two tests were carried out employing the architecture in Fig. 3.3(b), using the PPS generated by the WR ZEN in the first test, thus named as “WR ZEN” and the one coming from the cRIO-WR module, referred to as “WR Module”. For each of these cases, after the FPGA clock was completely locked and no transients appeared, the offset of the FPGA clock has been acquired for 2 hours.

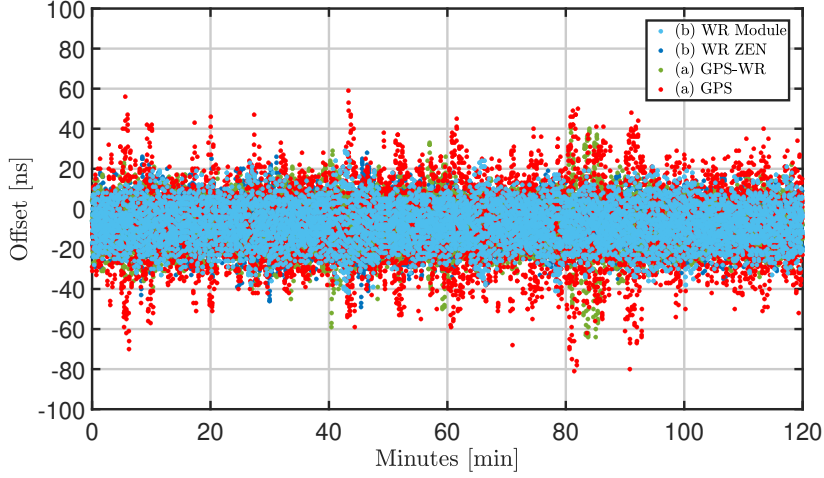


Figure 3.5: Offset measurements as function of time in the considered cases.

Fig. 3.5 shows the evolution of the offset with respect to the time reference of the FPGA clock (hereafter referred to as offset for brevity) as a function of time, for each of the performed tests¹, where the red dots represent the GPS case, the green dots the GPS-WR case, the blue dots the WR ZEN case, and the light blue dots the WR Module case. This color legend will be used for all the figures in this section. As a first comment, it can be noticed that none of the four configurations exhibit drifts or repetitive events, such as beats or modulations. It also appears quite evident that the average offset is not zero but consistent in the four cases, around -8 ns. This average is maintained constant over time and therefore it represents a systematic error introduced by the system. This may be caused by the delay introduced in the communication between the FPGA, the board, and the adopted Timekeeper library. It is probably due to the delay that the FPGA and or the boards used (NI 9467 and NI 9401, depending on the test considered) introduce in the reception of the PPS and consequently in the offset estimation. Obviously, being a practically constant value, it can be compensated for real-world applications, where time tagging techniques are needed (such as in synchrophasor estimation).

Moreover, it can be seen that the GPS and GPS-WR cases have the widest range of variation when compared to the others. Especially the former is affected by peaks around 60 ns and -80 ns. The WR ZEN and WR Module cases, on the other hand, do not exhibit these fluctuations and the data appear less spread. This is also confirmed by Fig. 3.6, which shows the histograms of the same four cases as in Fig. 3.5. The GPS synchronization has the largest offset variation

¹Please note that the data are superimposed for presentation purpose, but they are not related to the same time instants.

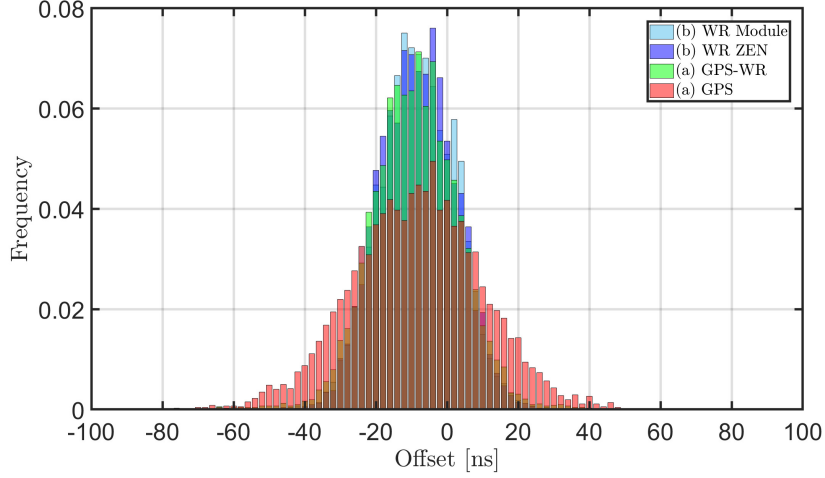


Figure 3.6: Histogram of the offset measurements in the considered cases.

with respect to the others, while no significant differences are visible for the other three cases. In addition, each test is characterized by a nearly Gaussian distribution.

To understand the statistical distribution of the data, the graphs in Fig. 3.7 show the quantile-quantile plots of the offset for the tests conducted at Cagliari on the left and at METAS on the right. For the tests conducted at Cagliari using GPS, it can be stated that the data appear almost Gaussian up to two standard normal quantiles and beyond they deviate from the ideal Gaussian assumption (dashed red line). In addition, the fits of the data have different slopes due to the reduced range of variation of the GPS-WR case. This is due to the different synchronization accuracy of the two GPS receivers used for the two tests (± 100 ns for the GPS case and less than ± 50 ns for the GPS-WR case). Considering the tests performed at METAS, the two distributions are almost coincident in terms of range. In the WR ZEN case, most of the samples lay on the ideal Gaussian, instead in the WR Module case the samples start to slightly deviate just before two standard deviations.

Table 3.2: Statistical Features of the Offset

Test	Offset [ns]			
	Max	Min	μ	σ
(a) GPS	59	-81	-8.00	18.30
(a) GPS-WR	40	-64	-8.32	12.12
(b) WR ZEN	28	-49	-7.88	10.52
(b) WR Module	29	-42	-7.85	10.50

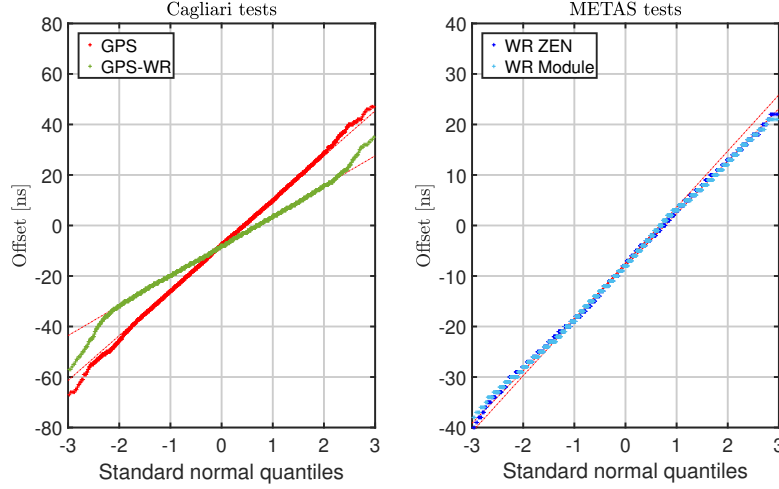


Figure 3.7: Quantile-quantile plot of the offset measurements in the four considered case.

The statistical features of the offset in terms of maximum, minimum, mean μ and standard deviation σ are shown in Table 3.2. It can be seen from both the graph and the table that the means do not vary appreciably and that the standard deviations are significantly smaller through the tests obtained at METAS using the WR protocol. As expected by observing the histograms in Fig. 3.6, the GPS case has the highest offset standard deviations. The GPS-WR case, on the other hand, is close in terms of σ to that obtained in the METAS lab, but it should be specified that the GPS-WR test was carried out using a more expensive high-performance GPS receiver if compared to the NI board. Nevertheless, the WR protocol yields the best results in terms of uncertainty and there is still margin of improvement, since, due to the 40 MHz maximum frequency and the jitter of the clocks of the two FPGAs, it is not possible to take full advantage of the sub-nanosecond accuracy achievable through WR.

To verify the stability of the synchronization obtained in the four cases, the fully overlapped Allan deviation [159] was calculated as a function of integration time and is reported in Fig. 3.8. Please refer to Appendix for a more in-depth discussion of the importance of this statistical index and explanation of its computation. The integration time has been varied from 1 to 2000 s in order to evaluate the temporal uncertainty over different time intervals. In this way, it is possible to compare the short- and long-term stability of the considered time propagation technologies. Consistently with the previous results, it can be seen that the cases tested at METAS using WR are those characterized by the lowest variability, where the Allan deviation goes, for example in the WR Module case, from $1.6 \cdot 10^{-8}$ to $8.5 \cdot 10^{-12}$. For the same integration time, in the GPS case

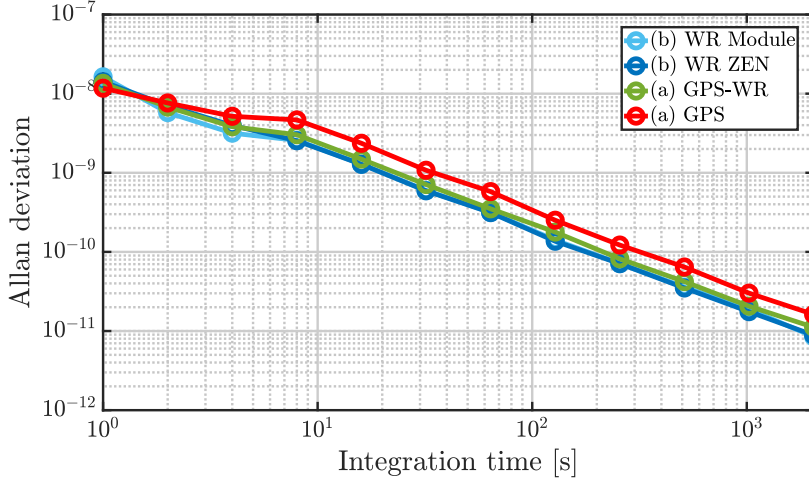


Figure 3.8: Allan deviation of the offset measurements as function of integration time in the considered cases.

values range from $1.2 \cdot 10^{-8}$ to $1.6 \cdot 10^{-11}$.

3.4 Examples of the contribution to measurement error of synchronization in power system applications

As mentioned in the previous section, synchronization represents a key technology to unlock new power system applications and enhance existing monitoring and control tools. For this reason, in this section, the implications of WR synchronization in two relevant measurement applications are discussed.

3.4.1 Harmonic synchrophasor measurement

The widespread diffusion of nonlinear devices has boosted the importance of harmonic monitoring in power systems. Moreover, accurately synchronized harmonic phasor measurements would enable studying the propagation of the disturbances, as well as detecting the most polluting users.

Synchrophasor estimates are performed and timestamped according to a UTC-synchronized time coordinate $\hat{t}$, subject to synchronization error. For the sake of simplicity, considering a rather small time interval between consecutive synchronization instants, it can be assumed that $\hat{t}$ results from an affine transformation of the actual UTC timescale t , hence

$$t = t_0 + (1 + \delta)\hat{t}. \quad (3.10)$$

Considering this model, $\hat{t} = 0$ is supposed to be the last synchronization instant (e.g., the second rollover) and t_0 is the corresponding offset. δ is the drift slope within the second.

Many popular methods for synchrophasor estimation are based on the so-called demodulation and filtering approach. It is the reference processing model adopted by [118], and it can be easily extended to harmonics measurements. Considering a power system having rated angular frequency ω_0 , operating under periodic steady-state conditions characterized by fundamental angular frequency ω_1 , the time evolution of the generic k -th order harmonic synchrophasor is thus

$$\bar{X}_k(t) = \bar{X}_k(t_r) e^{jk\Delta\omega(t-t_r)} \quad (3.11)$$

where t_r is the timestamp and $\Delta\omega = \omega_1 - \omega_0$. Samples are collected in the time instants $\hat{t} = t_r + nT_s$, with $n \in \{-N_b, \dots, N_e\}$, being T_s the sampling interval. The k -th order harmonic synchrophasor is extracted multiplying the samples by a complex exponential having angular speed $-k\omega_0$, and convolving the result with a proper Finite Impulse Response (FIR) low-pass filter whose coefficients are $w[n]$. Supposing that the frequency response of $w[n]$ and the small deviation between ω_1 and ω_0 result in negligible spectral interference, the synchrophasor estimate is

$$\hat{\bar{X}}_k(t_r) \approx \sum_{n=-N_b}^{N_e} w[n] \bar{X}_k(t_r) e^{jk\Delta\omega nT_s} e^{jk\omega_1(t_0 + \delta(t_r + nT_s))}. \quad (3.12)$$

The target is quantifying the accuracy degradation produced by synchronization error. For this purpose, the complex-valued error is defined as

$$\bar{\gamma}_k(t_r) = \frac{\hat{\bar{X}}_k(t_r) - \hat{\bar{X}}'_k(t_r)}{\hat{\bar{X}}'_k(t_r)} \approx \frac{\hat{\bar{X}}_k(t_r) - \hat{\bar{X}}'_k(t_r)}{\bar{X}_k(t_r)} \quad (3.13)$$

having introduced $\hat{\bar{X}}'_k(t_r)$ as the k -th order harmonic synchrophasor estimate under ideal synchronization, whose expression is given by (3.12) with $\delta = 0$ and $t_0 = 0$. It is worth noting that the real and imaginary parts of (3.13) virtually correspond to the magnitude and phase-angle error contributions induced by synchronization. With some substitutions, assuming that $w[n]$ is close to be symmetric with respect to $n = 0$, it can be obtained

$$\varepsilon_k(t_r) = \Re\{\bar{\gamma}_k(t_r)\} \approx -k^2 \delta T_s^2 \omega_1 \Delta\omega \sum_{n=-N_b}^{N_e} n^2 w[n] \quad (3.14)$$

$$\varphi_k(t_r) = \Im\{\bar{\gamma}_k(t_r)\} \approx k\omega_1(t_0 + \delta t_r). \quad (3.15)$$

The magnitude error $\varepsilon_k(t_r)$ represents the additional scalloping loss due to the non-ideal synchronization; it is generally negligible, in particular as far as $w[n]$ corresponds to a flat-top window. Fig. 3.9 shows the boxplot representation (95 % box and 99 % whiskers coverage) of the phase-angle errors in (3.15) com-

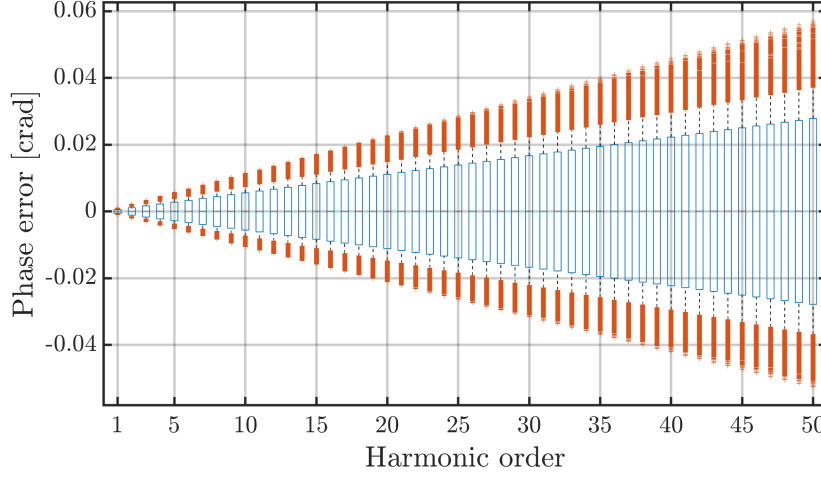


Figure 3.9: Boxplot of the phase-angle error due to synchronization as a function of the harmonic order for the WR Module case.

puted from the offsets found in the WR Module test ($\omega_1 = \omega_0 = 2\pi 50 \text{ rad/s}$). Interestingly, errors are at least an order of magnitude lower than 1 % TVE, thus enabling, in principle, an accurate harmonic synchrophasor measurement for all the harmonic orders of interest.

3.4.2 Traveling wave-based fault location

At fault occurrence on transmission lines, a high-frequency variation of voltage (traveling wave, TW) is generated and propagates towards both ends of the line in opposite directions. Such variation can be represented as a pulse that lasts in the order of microseconds or at most milliseconds, and its frequency content is mainly in the 2 kHz – 10 MHz range.

Without entering into considerations about applicability of TW-based fault location techniques, which should touch problems like identification of first traveling wave for correlation or location of fault close to terminals, it is evident that their performance depends pulse arrival time measurements, and thus on synchronization accuracy. In this regard, WR appears extremely promising and might represent a breakthrough.

For instance, when the double-ended TW method is adopted to locate faults, an accurate synchronization of instruments installed at both ends of a line (nodes 1 and 2, respectively) is a prerequisite. Considering two measurements of the arrival times of the TW referred to a common time reference, it is possible to express the two distances from the fault as:

$$\Delta x_1 = c \cdot (t_{a,1} - t_f) \quad (3.16)$$

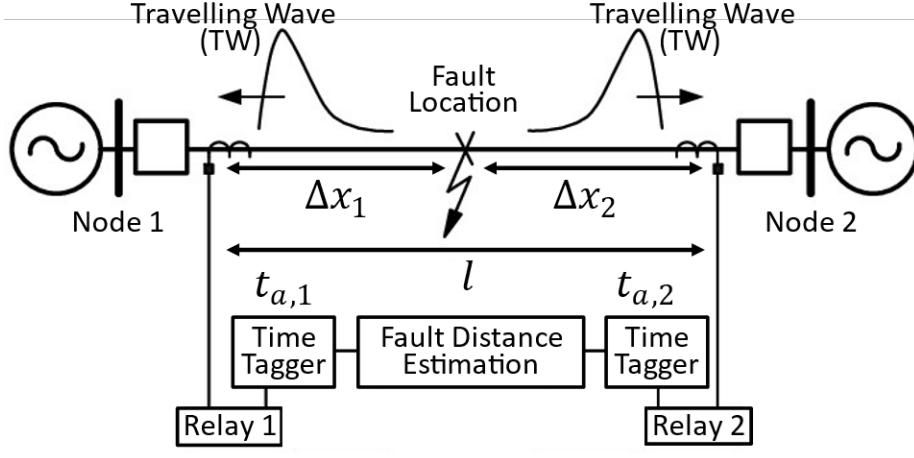


Figure 3.10: Block Diagram of a TW-based fault location system along a transmission line.

$$\Delta x_2 = c \cdot (t_{a,2} - t_f) \quad (3.17)$$

where c is the speed of the TW (which depends on distributed line capacitance and inductance and is close to the speed of light); $t_{a,1}$ and $t_{a,2}$ are the arrival time instants at node 1 and 2, respectively, while t_f is the unknown fault occurrence time instant. Combining the two expressions, they give

$$\Delta x_1 - \Delta x_2 = c \cdot (t_{a,1} - t_{a,2}) \quad (3.18)$$

If $\Delta x_1 + \Delta x_2 = l$ is the length of the line, substituting Δx_2 into the equation yields:

$$\Delta x_1 = \frac{l + c \cdot (t_{a,1} - t_{a,2})}{2} \quad (3.19)$$

To better clarify the concepts expressed, the diagram in Fig. 3.10 illustrates an example of a TW-base fault location estimation system along a transmission line.

An estimate of the fault point is thus obtained by measuring the two arrival times and using the formula in Eq. (3.19). Its standard uncertainty depends on the timestamping accuracy as follows:

$$\sigma_{\Delta x_1} = \frac{c}{2} \sqrt{\sigma_{t_{a,1}}^2 + \sigma_{t_{a,2}}^2} > \frac{c}{\sqrt{2}} \sigma_{\text{synch}} \quad (3.20)$$

where σ_{synch} is the timebase error due to synchronization.

With a typical 1- μs accuracy, the estimation accuracy is in the order of hundreds of meters. With dedicated and accurate GNSS receivers, according to the results of Section 3.3, it is possible to achieve a maximum error (considering only the contribution due to synchronization) of about 22 m.

Resorting to the WR, the advantage is two-fold: no multiple GNSS receivers

at line ends are needed while propagating high-accuracy atomic time reference and a standard uncertainty that is about 2 m (maximum error < 11 m) can be achieved. These values are lower bounds for overall accuracy with the considered platform but they already highlight the practical importance of improving synchronization technology and indicate a promising direction.

Chapter 4

Time quality assessment in all-digital measurement chain

This chapter initially introduces fully digital substations and all-digital measurement chains based on Sampled Value (SV) streams produced by Stand-Alone Merging Units (SAMUs) and received by Phasor Measurement Units (PMUs) with digital input for real-time monitoring. Regarding this, the importance of accurate synchronization in SAMUs and the implications of synchronization loss in downstream devices such as the PMU are also discussed. After that, the chapter proposes a method to evaluate the quality of synchronization using PMUs with digital input. Finally, experimental results of the implemented method are presented, which demonstrate the sensitivity and effectiveness of this method in assessing the quality of synchronization, allowing the reliability of power system measurements to be improved in various operational scenarios.

4.1 Digital substation architecture

In order to address the challenges described in Chapter 1, the distributed measurement infrastructure of the Wide Area Monitoring, Protection, and Control (WAMPAC) system can be realized by interconnecting different devices in different locations, characterized by remarkable performance in terms of both accuracy and latency [160], [161]. A complete refurbishment of each grid substation is practically unfeasible and therefore one of the first challenges to solve is represented by the interoperability, i.e., the capability of interconnecting instruments from different manufacturers and with different specifications [162]. In this sense, the family of standards IEC 61850 [26] constitutes the framework for the development of a fully digital and inter-operable substation, where dif-

ferent Intelligent Electronic Devices (IEDs) communicate with each other using the same protocols and data formats [27]–[29].

As shown in Fig. 4.1, common practice in primary substations was to have instruments connected to the High Voltage System (HVS) that estimate voltages and currents parameters. Instead, in fully digital substations, the task of interfacing with HVS is reserved to specific devices called SAMUs. These instruments communicate with IEDs that process the data to compute measurements.

In particular, the IEC 61850-9-2 introduces the SV protocol for the transmission of the secondary output of Non-Conventional Instrument Transformer (NCIT) [30]. In the typical measurement setup, a SAMU converts the analog output of the NCIT into a digital stream of SVs and publishes them on the substation network. In this way, the SVs are available not only for storage and post-processing applications in the control room, but also for an online estimation process as performed by IEDs [28], PMUs [31], [32] or Power Quality Instruments (PQIs) [33], [34]. The recent literature has proposed several approaches for calibration of NCIT [163], [164] or related instrumentation [165]. Furthermore, synchronization aspects have been addressed [166], but not much attention has been paid to the possible repercussions on the operation of real power systems.

In this regard, it is worth noticing that both the publisher (e.g., the SAMU) and the subscriber (e.g., the PMU) are disciplined to a common time reference with different synchronization requirements, as dictated by the corresponding reference standards. To this end, there exist several possible configurations. In a fully digital substation, it is reasonable to expect that both devices are connected to the same Precision Time Protocol (PTP) clock and configured with the same profile (for instance, Power or Utility profile). If the SAMU supplies a complete timestamp (as defined in IEC TR 61850-90-5 [143]), the PMU can operate without the local clock input. Another possibility is to use the optional SV timestamp field as defined in IEC 61850-9-2. Nevertheless, in this transition period, the possibility of independent time sources cannot be discarded a priori.

From a normative point of view, the recent IEC/IEEE Std 60255-118-1 [118] includes the Annex E for synchrophasor measurements using SV inputs with specific performance requirements in terms of estimation accuracy and reporting latency. Following a conservative approach, the Annex proposes to consider the PMU measurements as invalid as soon as the SAMU loses its time reference. Nevertheless, the recent literature has proposed an alternative approach that would allow to maintain the continuity of PMU service, yet flagging the measurements as less reliable [32]. As defined in IEEE C37.118.2 [36] Standard, the PMU measurement data packet contains a 32-bit field FRACSEC consisting

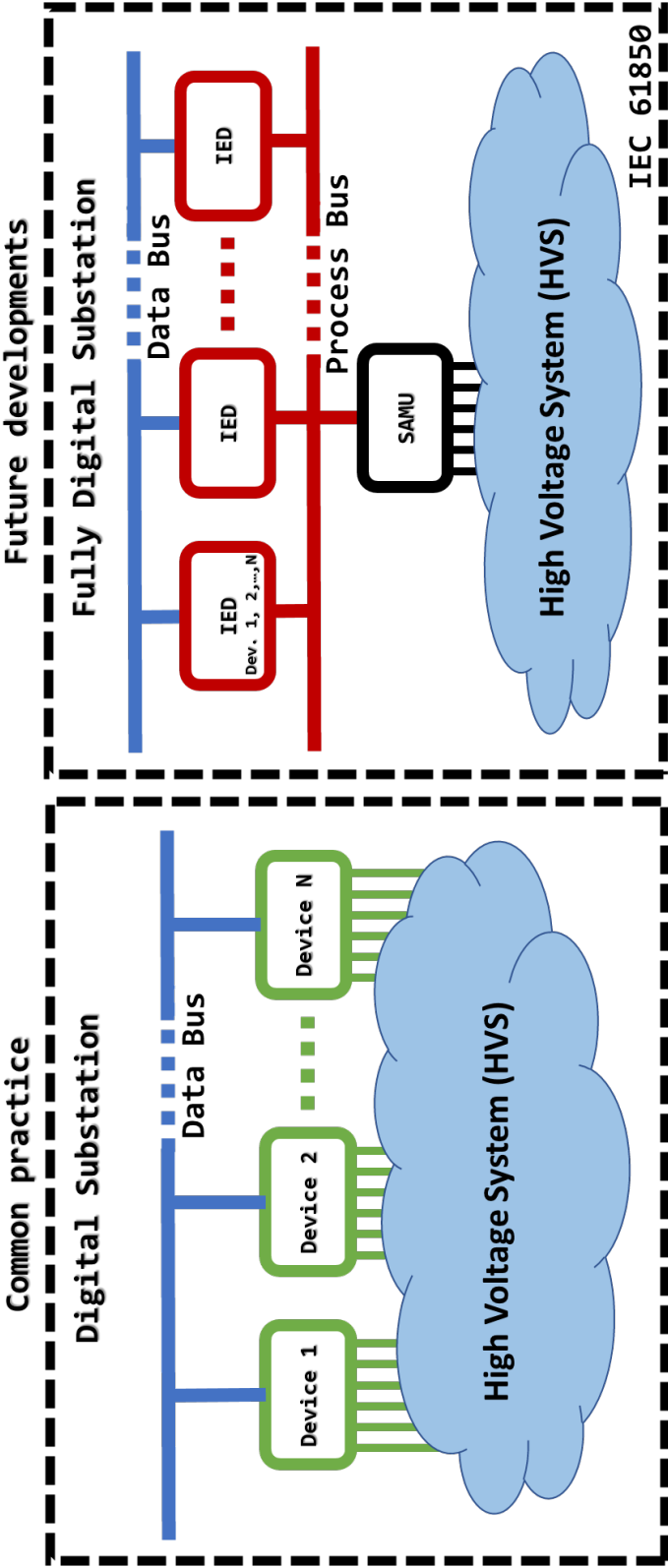


Figure 4.1: Comparison between architectures of traditional digital substation and future fully digital substation introduced by the standard IEC 61850.

of two components: a 24-bit unsigned integer for the fractional second, and an 8-bit message for the time quality flag. The latter accounts for the quality of the time information as well as the leap second status. In more detail, the time quality is expressed in seven classes, ranging from normal operation to clock failure. Each class is associated with a worst-case accuracy of the clock source, quantified in terms of deviation between the full second transition and the Coordinated Universal Time (UTC), e.g., the hexadecimal code 4 corresponds to a worst-case deviation of 1 μ s.

In this chapter, the case of a measurement chain where the SAMU and the PMU are disciplined by two independent clocks is considered. On the SAMU side, the temporary loss of the GPS antenna causes the absence of the time reference. Even after the restoration of the antenna connection, it is likely to expect several minutes are needed before the SAMU is actually locked to UTC. In the meantime, the SAMU keeps producing the SV stream, but the packets timestamp is flagged as unreliable. On the PMU side, despite the device is still locked to UTC, the PMU time quality shall be updated to reflect the lack of synchronization until the SAMU synchronization has been restored. As a consequence, a similar event may cause a service interruption lasting even tens of minutes [167]. On the other hand, modern PMUs rely on clocks characterized by remarkable performance in terms of short-term stability and resolution in the order of sub-millisecond scale [96], [168], [169].

4.2 Fully digital measurement chain

Modern substations are rapidly transitioning towards a fully digital and automated communication and control paradigm. The different IEDs installed in the substation do not operate independently and autonomously, but communicate with each other and with the control room in order to achieve a prompter and more efficient response to unexpected or even potentially dangerous events.

As already mention in the precedent section, IEC 61850 introduces the infrastructure and defines the message data types and performance requirements for the IED communications. In this regard, Fig. 4.2 presents the typical measurement chain.

The power signals of interest are acquired via a suitable instrument transformer (IT). A merging unit (MU) collects the digitized signals (e.g., three-phase voltage and current), and encloses them in a SV stream. The digital data stream is published over the substation network and made available to other measurement and control devices. For instance, a PMU may retrieve the SV stream and process it in order to produce the corresponding measurements of synchrophasor, frequency and rate of change of frequency (ROCOF).

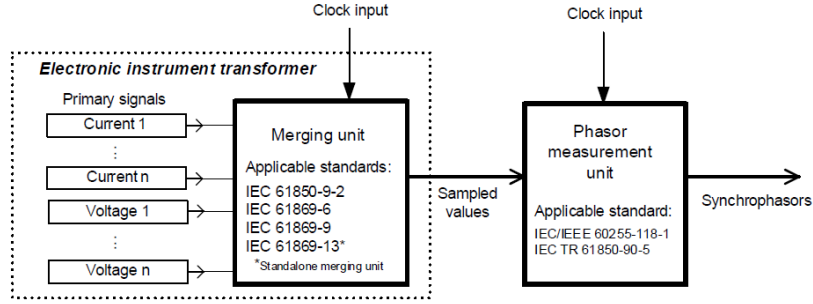


Figure 4.2: Measurement chain for synchrophasor measurements based on SV inputs, as per IEC/IEEE 60255-118-1 [118].

In the framework of IEC 61850, the communication relies on an Ethernet physical layer. It is thus reasonable to assume that also the synchronization signals could be transmitted over the same network, e.g. via PTP. A single PTP clock, aligned to UTC via a GPS antenna, could provide an accurate and stable synchronization to all the IEDs installed in the substation, with a remarkable advantage in terms of cabling reduction and implementation costs. In order to address the challenging requirements typical of power system applications, a specific PTP profile is defined in the IEC 61850-9-3. The so-called Utility profile introduces stringent requirements for the grandmaster, transparent, and boundary clocks. The objective is to guarantee an end-user timing accuracy not exceeding $1\text{ }\mu\text{s}$ even in a network configuration consisting of 15 transparent clocks and/or 3 boundary clocks. In the same profile, a specific field clockClass accounts for the synchronization status and allows for discriminating between normal operations, loss and recovery of the synchronization. In particular, during the restoration after a synchronization loss, the clock status is progressively updated based on the expected time error: holdover, $1\text{ }\mu\text{s}$, 250 ns, recovered.

Indeed, it is important to underline that synchronization status has a crucial role for PMU measurements. Most PMUs use the external time-source signal not only to define the full second transition and thus the measurement timestamp, but also to discipline the internal clock. As a consequence, an erroneous or a missing synchronization may affect not only the reporting time reference and thus the correct aggregation with other PMU measurements, but also the estimated quantities, particularly the phase angle and its derivatives (i.e. frequency and ROCOF).

In this regard, the PMU Std [118] defines a performance target for PMUs in normal operating conditions of a Total Vector Error (TVE) not larger than 1 %. In the absence of magnitude errors, this corresponds to a phase angle error of 10 mrad. At a nominal system rate of 50 Hz, such a phase angle is equivalent

to a timing error of nearly $32\ \mu\text{s}$. This explains the importance of accurately determining the PMU time quality: values over $10\ \mu\text{s}$ could be associated with a measurement chain that is prone to large TVEs.

Based on these considerations, the configuration in Fig. 4.2 provides significant advantages. First of all, the adoption of a predefined and common data format guarantees a full-interoperability between any SAMU or PMU compliant with the pertinent standards. Moreover, the possibility of separate (and possibly independent) time sources for SAMUs and PMUs guarantees a redundant and effective control of the synchronization state of the two devices. In this regard, it is worth noticing that both SAMU and PMU packets include indicators of the synchronization status and quality of the respective device.

At the SAMU level, the synchronization signal triggers the publishing of the SV packet with packet count 0, that shall correspond to the full second transition. Each SV packet contains a SmpSync field that denotes the status of the MU synchronization. More precisely, the MU can be classified as synchronized or not. In the first case, the device discriminates three clock categories: global area, local area and unspecified local area. The lower the quality of the clock, the higher the uncertainty propagated to the following elements in the measurement chain, i.e. the PMU.

In a PMU with digital inputs (briefly, SV-PMU), the synchronization signal can be used to timestamp the measurement data packets but also to check the consistency of the SV input stream (e.g., with a check on the SV packet latency). The PMU measurement data packets include several fields related to the synchronization status and quality. In the STAT field, the PMU Sync Error bit is set to 1 in the absence of a valid external time synchronization (e.g., loss of satellite tracking, time input connection failure, or time source no more locked to UTC). In the FRACSEC field, instead, 32 bits are organized in two words. The first 24 bits represent the time elapsed since the last UTC full second. The last 8 bits include the PMU Time Quality (TQ) flag, called precisely PMU_TQ. More precisely, 3 bits express the maximum time error with respect to UTC ranging in seven different levels from 100 ns to 10 ms. Finally, other two bits are used to quantify the Unlock Time, i.e. the count of seconds elapsed since the last loss of synchronization was detected.

Based on all these information, it is possible to accurately track the synchronization status of the entire measurement chain and promptly detect any fault or service interruption. In this sense, it is worth noticing that the PMU Std adopts a quite conservative approach. In the presence of a synchronization loss, the PMU measurements are flagged as unreliable until the synchronization source has returned to normal steady-state operations. In a hierarchical perspective, it is necessary that the PTP clock first, and the SAMU second,

restore their time source integrity before the SV-PMU can resume its normal operation. In GPS-based systems, this may take up to few tens of minutes, causing significant restrictions in many grid services as state estimation, fault location and power flow control.

On the other hand, modern PMUs rely on high-performing industrial controllers, typically equipped with real-time (RT) operating systems. In most cases, the internal clock is characterized by remarkable performances in terms of short-term stability and timestamp accuracy.

4.3 Proposal of Sampled Value PMU with time quality evaluation functionality

Assuming that a SAMU relies on the synchronization source to periodically adjust the publication period of SV stream, in case SAMU has variations in synchronization accuracy, the publication period is heavily affected. For example, considering a generic Global Navigation Satellite System (GNSS) as the synchronization source for SAMU, the pulse per second (PPS) received from the GNSS receiver will allow SAMU to correct its synchronization offset every second, periodically adjusting the start of the SV stream and thus varying the publication period. The difference between the timestamp of the SV packet corresponding to PPS rising edge (thus with count index equal to zero) and the timestamp of the previous packet can be defined as $\hat{t}_p$.

Since the SAMU relies on the PPS to periodically adjust the start of the SV stream, a consistent synchronization offset, e.g., due to a loss of synchronization, leads to a time period offset t_p^{os} , which is defined as the difference between $\hat{t}_p$ and the nominal publishing period t_{p0} as follows:

$$t_p^{os} = \hat{t}_p - t_{p0}. \quad (4.1)$$

By suitably timestamping the incoming SV packets using a dedicated synchronization source, the SV-PMU can evaluate $\hat{t}_p$ and consequently can compute t_p^{os} with Eq. (4.1), assessing in this way the deviation of the SAMU full second transition from the UTC one. In this way, the SV-PMU can check the consistency of the SV publishing period and verify indirectly the synchronization accuracy of the SAMU. This value can be considered as a rough estimation of the deviation of the SAMU clock with respect to UTC. By properly tracking the oscillations of this value, the SV-PMU is able to update its time quality flag PMU_TQ autonomously at each second.

To update the PMU_TQ, a dedicated algorithm, summarized as a pseudocode in Algorithm 1, has been proposed in addition to the *Decoder* part of

CHAPTER 4. TIME QUALITY ASSESSMENT IN ALL-DIGITAL MEASUREMENT CHAIN

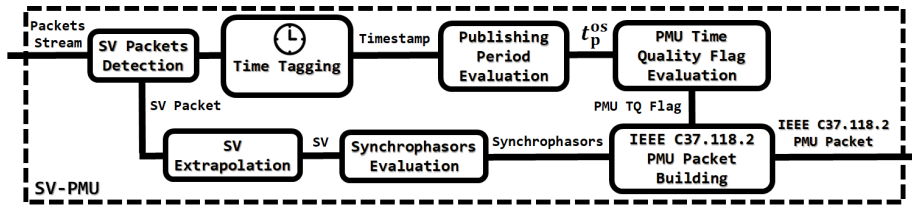


Figure 4.3: Block diagram of proposed SV-PMU with time quality evaluation functionality.

the SV-PMU software structure of [32]. In the *Decoder* part of the SV-PMU software, as can be seen from the block diagram in Fig. 4.3, when the SV packet belonging to the stream of interest is identified, the method performs two actions in parallel. On the one side, it extracts the sample count and the measured current and voltage values, and, on the other side, the arrival time of the SV packet is obtained in terms of an absolute timestamp provided by the synchronization source. The proposed time quality evaluation algorithm is added as a new parallel task inside the *Decoder* while loop.

In the following description, the main lines of the code that updates the PMU_TQ flag are described. In the first if-statement, t_p^{os} and its previous value (note that in the Algorithm 1 the apices ' and ' refer respectively to the previous value and the value before previous of the variable) are analyzed and the current time quality value tq_v is then set according to the following three different rules: either the absolute value of t_p^{os} is the highest value reached, or t_p^{os} is detected as a peak, or even the previous t_p^{os} value is detected as an outlier peak. In the first two cases, tq_v takes the absolute value of t_p^{os} , whereas in the third, it is set to its next-to-last value. If none of these three conditions is satisfied, tq_v is set to its previous value.

It is important to note that, in the Algorithm 1, the peak corresponds to a significant positive or negative deviation in the measured t_p^{os} , corresponding to a proportional deviation of the SAMU synchronization offset.

Subsequently, the value of tq_v is checked. If it is below the chosen time quality threshold TQ, the counter increases, otherwise the count is set to 0. This conservative condition implies that the value of tq_v is required to fall below the TQ threshold for a specified number of consecutive seconds before updating the status of the PMU_TQ flag. In this thesis, the fulfilment of this condition requires the counter to exceed the threshold for five consecutive iterations, i.e. five seconds. This value has been chosen on the basis of empirical observations of re-synchronization curves in order to avoid the undesirable impact of certain oscillations that can be misidentified by the algorithm.

This proposal does not guarantee the accuracy of the PMU measurements

Algorithm 1: TQ-evaluation in RT

```

Input  :  $t_p^{os}$ 
Output: PMU_TQ
1 Init :
2   count  $\leftarrow$  0;
3   tq_v'  $\leftarrow$  max_v  $\leftarrow$   $t_p^{os}$ ;
4 while SV stream detected do
5   if  $|t_p^{os}| > \text{max\_v}$  then
6     max_v  $\leftarrow$   $|t_p^{os}|$ ;           /* update max value */
7     tq_v  $\leftarrow$  max_v;         /* set to max value */
8   else if peak is detected then
9     tq_v  $\leftarrow$   $|t_p^{os}|$ ;       /* set to absolute value of peak */
10  else if previous value is an outlier peak then
11    tq_v  $\leftarrow$  tq_v'';        /* set to value before previous */
12  else
13    tq_v  $\leftarrow$  tq_v';         /* set to previous value */
14  end
15  if tq_v < TQ then           /* check TQ threshold */
16    count  $\leftarrow$  count + 1;
17  else
18    count  $\leftarrow$  0;
19  end
20  if count > 5 then
21    update PMU_TQ;           /* update TQ flag */
22  end
23 end

```

but provides a preliminary assessment of the measurement chain consistency. Instead of waiting the full restoration of the GNSS signal for the SAMU, the control room can consider the PMU measurements and weight their reliability with a quantitative index.

4.4 Implementation of the measurement setup

This section describes the measurement setup used to characterize the proposed method in plausible operating conditions.

In view of a full observability and controllability of the different synchronization and processing stages, no commercial SAMUs or PMUs have been employed. Conversely, in order to perform tests in the presence of synchronization issues, a modular measurement architecture, composed of industrial real-time controllers equipped with a Field Programmable Gate Array (FPGA) chip and expansion boards is taken into account analogously to that used in Chapter 3. Specifically, the SAMU and the SV-PMU functionalities are implemented

in two industrial controllers, namely a National Instruments CompactRIO-9040 (NI cRIO-9040) [170] and a NI cRIO-9068 [158], as shown in Fig. 4.4.

In particular, the first device, implemented with the NI cRIO-9040, has been developed to replicate the behavior of a commercial SAMU with the additional functionality of simulating the acquisition of SVs and the packets transmission conforming to the IEC 61850-9-2 standard [30].

The SAMU encodes and sends out a series of SV packets, using the features of specific functions included within an *ad hoc* Shared Object library. This library is an extended version of the one presented in [32] with additional functionalities dedicated to the publishing of SV packets in RT. In this context, the function designed to send the packets out of an Ethernet adapter is called once for each SV within the run of a timed loop. The execution frequency of such a loop is regulated by the 10 MHz internal clock of cRIO. By properly programming the SAMU, it is possible to control the SV publishing process guaranteeing a sufficiently accurate and stable publishing period (typically in the order of hundreds microseconds).

As soon as a new full second transition is detected, the SAMU starts its SV data stream with a publishing period that can be either stationary or time-varying depending on the variation of synchronization accuracy. Only in the case of the tests in Section 4.5.1, for the purpose of exhaustively testing the proposed SV-PMU, the publication period is set as stationary with specific nominal publication periods or variable with a sinusoidal law and thus does not depend on changes in synchronization accuracy. The SV publishing period has been defined with a quantization level of 1 μ s, but a finer resolution would be possible by exploiting the FPGA board of the same controller.

The second device, implemented with the NI cRIO-9068, allows to emulate the behavior of the proposed SV-PMU explained in detail in the precedent section. For the implementation, the NI cRIO-9068 includes a packet sniffer functionality that allows for capturing and decoding only the SV packets of interest. Each packet is timestamped according to the internal time-base and the time difference between consecutive packets is computed. When the received SV packet corresponds to the transition of the second given by the PPS rising edge, $\hat{t}_p$ is calculated and subsequently, based on the nominal SAMU publication period t_{p0} , t_p^{os} is computed using Eq. (4.1). After that using the Algorithm 1, the time quality of SAMU is evaluated and its time quality flag PMU_TQ is updated independently, updating it into the IEEE C37.118.2 packet. Further details about the realization and characteristics of the SV-PMU are available in [32].

Precisely, the NI cRIO-9040 is equipped with a 1.30 GHz Dual-Core Intel Atom E3930 CPU, 2 GB DRAM and a Xilinx Kintex-7 7K70T FPGA charac-

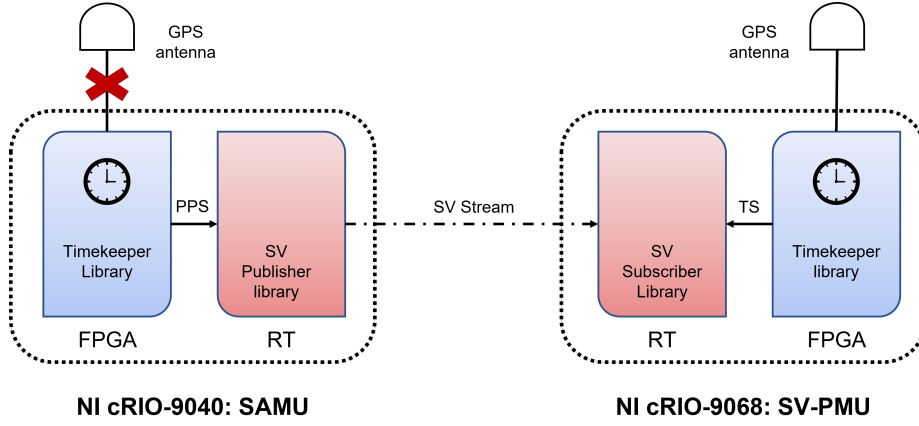


Figure 4.4: Measurement setup for reproducing the synchronization loss at SAMU level.

terized by a maximum time-base accuracy of 200 ppm and typical of 40 ppm at 25°C. Instead, the NI cRIO-9068 is exactly the same as the one used in Chapter 3 and thus it will not be discussed further here. Each controller is equipped with high-performance processing units running Linux RT operating systems, and a GPS receiver module (NI 9467 as in Chapter 3) for locking their internal clock to UTC. By means of a short point-to-point connection, the SV packets output by the SAMU are directly transmitted to the PMU. In this sense, the delay introduced by the two Ethernet adapters and the cable can be considered negligible. The NI 9467 boards provide the synchronized PPS signal to the FPGA and, through the dedicated libraries FPGA Timekeeper API, already mention in Chapter 3, allow the different clocks of the prototyping devices to be tuned. In free running, a typical NI cRIO real-time controller clock has a drift of ± 40 ppm, and the FPGA module clock has a drift of ± 100 ppm, which would cause time drifts up to several seconds per day if not synchronized.

As in the software presented in Fig. 3.4 of Chapter 3, the PPS signal is considered to discipline the FPGA module, allowing the time tag to be executed up to each FPGA clock tick (25 ns).

To replicate possible synchronization problems encountered in commercial SAMUs equipped with internal GPS receiver, synchronizing signal loss and recovery events are realized in a laboratory environment. In fact, by acting on the GPS antenna connected to the NI 9467 board, it is possible to simulate the absence of the synchronizing signal. In a real context, the lack of the synchronizing signal can be caused by various factors, such as the momentary presence of an obstacle that interferes with the visibility of the satellites or by intentional actions to degrade the synchronizing signal. As shown in Fig. 4.4, the

effect of disconnection of the synchronizing signal causes a momentary absence of the signal that disciplines the FPGA clock, causing a time-dependent drift. When the synchronizing signal is restored, the clock of the FPGA changes from a free-running state to a disciplined state.

However, during the restoration of the synchronizing signal, the FPGA clock takes a variable time to re-establish synchronization levels acceptable to satisfy requirements for SV-based applications. The re-establishment time depends on hardware and software factors such as the type of clock employed in the FPGA and the libraries for managing the synchronization discipline.

4.5 Test and results of the implemented SV-PMU

This section presents the tests and results obtained for the implemented SV-PMU by appropriately evaluating the sensitivity of the SV packet timestamping functionality under realistic conditions with stationary and variable publishing period and finally assessing the time quality estimation of the proposed algorithm.

It is worth mentioning, that in the following presented results, without loss of the generality, a nominal publishing period of 250 μ s is considered. In this case, each packet contains the information of a single SV, or more technically: the number of Application Service Data Units (ASDUs) per frame is equal to 1 (see [30]). This means therefore that the SVs are sent with a configuration whereby the publishing rate is equivalent to the sampling rate, namely 4 kHz. Similar results can be achieved, by considering also the other sampling and publishing rate configurations listed in the IEC 61869-9 Standard [171].

4.5.1 Time synchronization sensitivity of SV-PMU

Before measuring the periods between the SV published by the SAMU, the accuracy of the proposed SV-PMU achieved with the RT controller needs to be assessed.

For this reason, the sensitivity of the SV-PMU clock oscillator is evaluated using two different measurement approaches. The first method consists of timestamping a SV stream published with a stable and constant publishing rate, which reproduces the normal operation of a synchronized SAMU. The second method, instead, evaluates the timestamping functionality when the SV stream is sent with a non-constant publishing period between packets. This condition may occur, for example, if there are: synchronization losses in the reference time system, internal malfunctions of the publisher or some issues in the communication system.

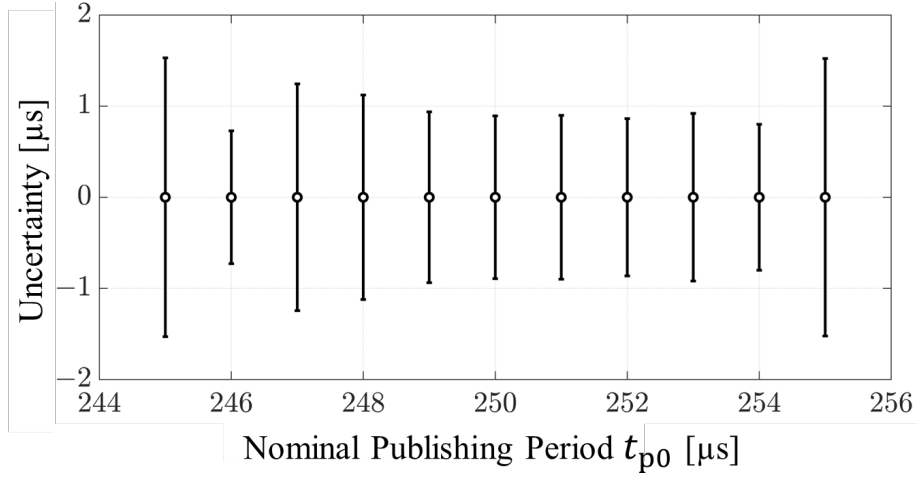


Figure 4.5: Symmetric uncertainty ($k = 2$) affecting the SV-PMU publishing periods measurements as a function of nominal publishing periods t_{p0} ranging from 245 to 255 μs .

It is important to note that, in the following test, in order to have a consistent number of measurements in a reasonable short time, the measurement of not only the publishing period corresponding to the PPS rising edge $\hat{t}_p$ but those of each SV packets received has been used.

Fig. 4.5 shows the symmetric uncertainties affecting the time periods elapsing between the SV packets as measured by the SV-PMU. Each of these uncertainties corresponds to two standard deviations determined from the values, computed by taking the difference between 120000 timestamped SV packets, normally distributed around the corresponding nominal publishing period t_{p0} . For this uncertainty evaluation, eleven nominal publishing periods have been chosen within the interval ranging between 245 and 255 μs , whose limits represent a 5 μs deviation from 250 μs , which corresponds to a common publishing period adopted by many commercial SAMUs.

There are two main considerations that can be drawn from the results presented in Fig. 4.5: first, 95% of the computed publishing periods lie within the 1 μs margin, or slightly above, for each nominal publishing period; second, no trend for uncertainty as a function of nominal publishing period is visible. This is important because it allows assuming that the uncertainty settles on the 1 μs value, which represents the resolution given by the instrument clock, regardless of the measured publishing period.

In order to characterize the reactive response of the proposed SV-PMU in computing the timestamp difference between packets from a SV stream, whose publishing rate is not constant, sinusoidal variations have been considered. In

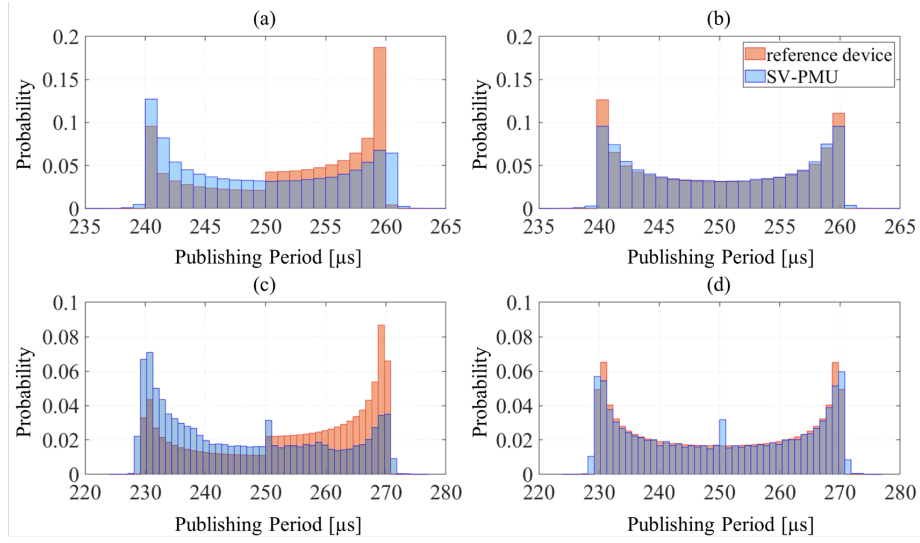


Figure 4.6: Distribution of published (orange) and received (blue) SV, with a sinusoidal deviation w.r.t. nominal publishing period of 250 μs realized with the following amplitude and frequency parameters: (a) $A = 10 \mu\text{s}$ and $f = 6 \text{ Hz}$, (b) $A = 10 \mu\text{s}$ and $f = 4 \text{ Hz}$, (c) $A = 20 \mu\text{s}$ and $f = 6 \text{ Hz}$, (d) $A = 20 \mu\text{s}$ and $f = 4 \text{ Hz}$.

this way, it is possible to continuously measure all the SV timestamp differences within the defined range of values that deviate from the nominal publishing period t_{p0} . Furthermore, a sinusoidal function allows to test the flexibility of the receiver clock by varying: the amplitude (A), corresponding to the maximal deviation from the nominal publishing period t_{p0} , and the frequency (f) at which this deviation varies.

In this test, the SAMU timestamps the SV packet output and computes the difference with the timestamp of the previous SV packet output in a similar way to the SV-PMU. For this reason, the SAMU can be considered the reference in this test since the variation in the publishing period of SV packets is tuned and does not depend on the variation in synchronization accuracy (that was constant for all the test duration).

Fig. 4.6 compares the distributions of 120000 publishing periods measured between packets sent by the reference SAMU and packets received by the SV-PMU. The time periods between the published SV deviate from the nominal publishing period of 250 μs according to a sinusoidal function. Four characterization measurements have been performed with sinusoidal deviations realized by combining two different amplitudes A of 10 and 20 μs and frequencies f of 4 and 6 Hz. A comparison of the histograms suggests that the SV-PMU appears to be more affected by the increase in frequency than the increase in amplitude of

the publishing rate variation. The SV-PMU distributions shown in Fig. 4.6 (b) and (d) in fact reproduce better the distributions calculated with the values of the reference SAMU than in the case shown in Fig. 4.6 (a) and (c).

In all the considered configurations, the SV timestamping functionality does not lose a noticeable number of packets, neither it does introduce outliers. Indeed, the recovered distributions are extremely similar to the original ones, as measured by the reference device. In this sense, it is reasonable to say that an analysis of the distribution of the SV timestamps would allow to detect oscillatory publishing rates and thus trigger suitable control actions and checks on the SAMU.

The statistic results presented in Fig. 4.6 demonstrate the reliability of the SV-PMU in measuring the periods between SVs not published with a constant rate. The limits of 4 Hz for the frequency at which the packets can be correctly timestamped by the device is far beyond the values needed for the measurements presented in this chapter.

4.5.2 Results of SV-PMU time quality evaluation

This section describes an experimental validation of the proposed time quality evaluation in SV-PMU. This analysis aims to evaluate the actual performance achieved with the implemented SV-PMU in the presence of anomalies in the synchronization chain. More specifically, the case of synchronization failing due to disconnection of the GPS antenna from the NI 9467 board in the reference SAMU system for a few minutes is considered. Such a communication malfunction causes the interruption of the PPS signal output from the FPGA that represents the SAMU synchronization source (see Fig. 4.4). This leads to a value change in the SV field SmpSync, whose integer is set to 0 indicating an absence of synchronization signal. When the antenna is reconnected, the PPS signal is immediately resumed thus causing the value of the SmpSync field to change again considering the SAMU as synchronized. But, in the reality, it can be observed a synchronization offset between the FPGA clock and the synchronization source (GPS signal) due to the readjustment of the NI 9467 receiver.

In the considered measurement setup, it is not possible to have a mathematical model of the behavior of the FPGA clock disciplined by the NI 9467 GPS receiver due to the absence of specifications in the official datasheet. Nevertheless, based on the different datasets obtained during the experimental characterization, it is reasonable to assume that a synchronization loss induces a damped oscillation whose magnitude, frequency and damping parameter depend on the loss duration and other environmental factors [172].

As proof of this, it is possible to observe in Fig. 4.7 the synchronization offset

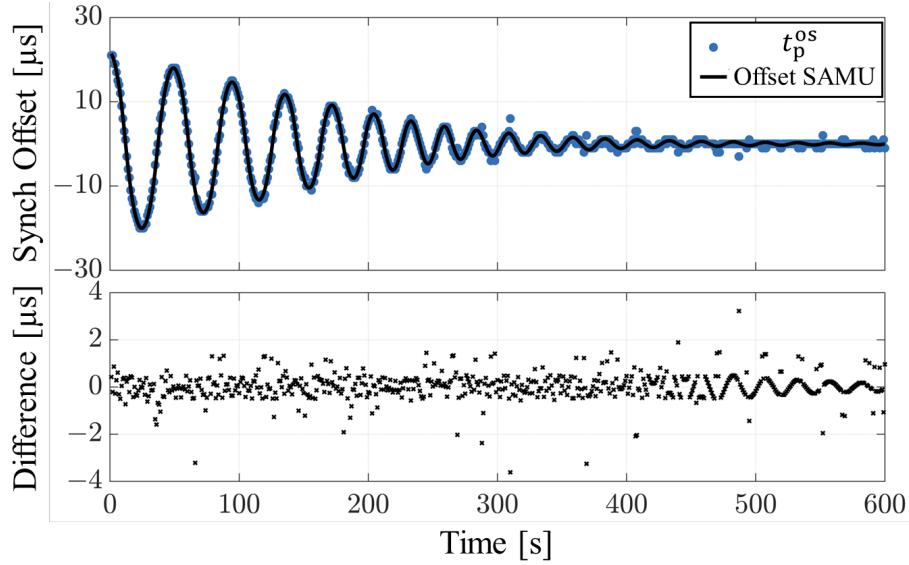


Figure 4.7: Top plot: synchronization offset of the FPGA clock measured by the NI 9467 synchronization board (solid line) compared against t_p^{os} as measured by the SV-PMU (dots) vs time. Bottom plot: difference between measurements.

of the FPGA clock obtained with the NI 9467 synchronization board (solid black curve). The synchronization offset is computed with a reporting rate given by the PPS, as seen in Chapter 3.

Fig. 4.7 also compares the synchronization offset of the SAMU with t_p^{os} (blue dots in the plot) as determined by the SV-PMU. For the test presented, the communication with the GPS antenna has been cut for 10 min and 20 s. It can be noticed that, at the beginning of the top plot of Fig. 4.7, thus as soon as the synchronization signal is restored in the SAMU, the FPGA clock of the SAMU is affected by an offset of 20 μ s from the GPS time. When the connection is resumed, the reference system oscillator is re-disciplined and, after ten minutes, the offset decreases to values that fall within the typical uncertainty of the device clock when locked.

The bottom plot of Fig. 4.7 reveals the reliability of the SV-PMU in detecting anomalies in the synchronization of the SAMU with respect to the reference time. The difference between the offset values measured by the reference system and the SV-PMU is comparable, for the most of cases, to the accuracy level of the tested instrument as characterized in Section 4.5.1. It is interesting to notice that the sinusoidal pattern, visible after about 480 s, occurs when the SAMU synchronization offset falls below the 1 μ s resolution of the SV-PMU.

Proven that the SV-PMU is able to measure the offset from the reference

time synchronization system of the SAMU with high accuracy, it is reasonable to attempt to provide the real-time estimation of the synchronization status through the update of the time quality PMU_TQ flag through the proposed Algorithm 1.

The results of the reliability of Algorithm 1 are illustrated in Fig. 4.8 using the example of reference clock synchronization malfunction previously presented. It can be appreciated how the time quality levels, evaluated at each full second transition, are consistent with the offsets t_p^{os} from the nominal publishing period t_{p0} corresponding to the detected peaks.

The sparse outliers occurring during the ten measured minutes modify the curve instantly, but do not affect the final result. This is because of the condition set at lines 11-12 of Algorithm 1, which prevents an outlier mistaken for a peak from representing a problem in terms of the time quality level reported by SV-PMU. Should this occur, during the subsequent iteration of the algorithm (i.e. the second after) it would be possible to detect this failure and revert it to the next-to-last valid value. Thus, these misdetected peaks only have the effect of reproducing isolated dips or spikes, whose value is not maintained for more than one second. Nevertheless, the real-time condition prevents these values from being removed.

The outlier of about $-10 \mu s$ occurring after 66 seconds in Fig. 4.8 is a good example to understand the contribution of this condition on the robustness of the algorithm. Without implementing it in fact, the current time quality variable tq_v would be set to the false peak value and kept for 6 seconds until the next true peak. Hence, as this value falls below the time quality threshold, it would lead to an improper update of the PMU_TQ flag.

By taking as an example the time quality threshold of $10 \mu s$ depicted in Fig. 4.8, it is worth noticing that the values evaluated by the algorithm fall below this line starting at 2 min 54 s from the beginning of the analysis while remaining there until the end. This means that the PMU_TQ flag contains useful information on the synchronization status seven minutes earlier than the reference clock system is completely re-locked to the GPS signal. Exploiting this additional information could therefore enable a faster comeback to quasi-normal operating conditions.

It is worth noticing that the proposed approach does not represent a metrologically rigorous assessment of SAMU synchronization. Indeed, the accuracy and reliability of the obtained results are strongly dependent on the precision and resolution of the PMU timestamping functionality. Nevertheless, given the remarkable performance of modern industrial controllers, the proposed method may represent a valuable alternative solution in the presence of synchronization losses. In fact, in any monitoring and control system, a partial yet not-fully

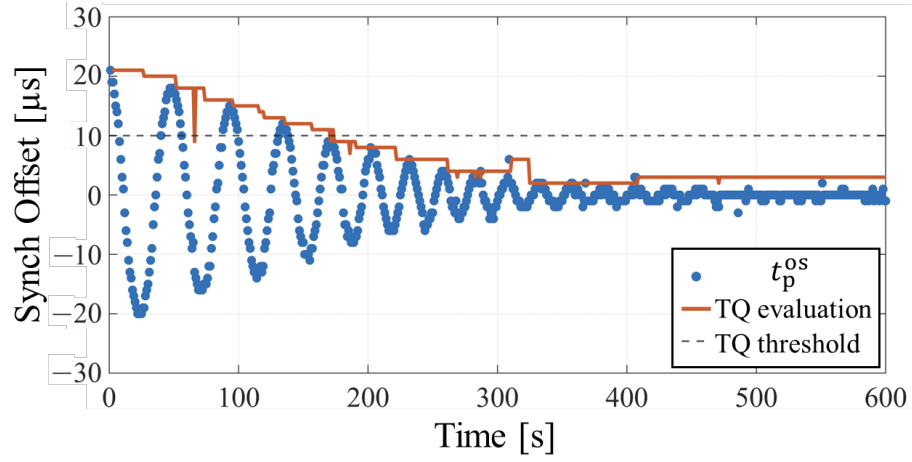


Figure 4.8: t_p^{os} as measured by the SV-PMU (dots) vs time and RT time quality evaluation algorithm values (solid line). Dashed line represents the time quality threshold of 10 μ s.

reliable observation is preferable to a complete absence of information.

Chapter 5

Latency measurement architectures for WAMPAC systems

Following the necessity of appropriately assessing the latency of Wide Area Monitoring, Protection, and Control (WAMPAC) systems, after introducing the various contributions to the overall WAMPAC latency, this chapter presents the design and implementation of a WAMPAC latency characterizer, i.e., an instrument to accurately measure latency in purely digital WAMPAC systems, which allows testing several Phasor Measurement Units (PMUs) at once, with different protocols, packet sizes, and reporting rates. A modular instrumentation platform by National Instruments (NI) equipped with real-time hardware is used to prototype the proposed design. In order to carry out a metrological verification of the functionality and performance of the proposed instrument, a second device has been designed and implemented to act as an emulator of a WAMPAC system, which is programmable and able to mimic real WAMPAC features. Then, the experimental characterization of a WAMPAC system based on software tools commonly used in real Wide Area Monitoring System (WAMS) and WAMPAC applications is illustrated and confirms the validity of the proposed measurement platform. Finally, a customized version of the instrument has been used to characterize the latency of a WAMPAC prototype of the Italian transmission system operator (TSO) based on 4-20 mA actuator technology, emulating 2 PMU Transport Control Protocol (TCP) data streams, with fixed reporting rate and packet size.

5.1 Latency challenges in WAMPAC systems

WAMPAC systems, as emphasized in Chapter 1, have emerged as a critical technology to improve the reliability, resilience, and stability of modern power grids. Since these systems are required to make rapid decisions and control actions on the grid, they are characterized by stringent time constraints.

In this context, one of the critical aspects in the design and implementation of WAMPAC systems is the total latency, i.e., the time delay between the occurrence of an event in the electric system and the execution of associated control actions. Latency can greatly affect overall system performance, with possible implications for the safety, reliability and stability of the grid. Therefore, it is useful to evaluate the latency of each element within the monitoring and control chain. In this regard, it would be beneficial to propagate time properly across the entire network infrastructure [173].

Depending on the implemented algorithm, PMUs can introduce considerable delay into WAMPAC systems and, for this reason, the literature has focused on their latency characteristics and requirements. In [174], for instance, a low-latency (less than 10 ms) algorithm, suitable for control applications, is proposed. Characterization of PMU latency is analyzed in [175], [176].

As seen in Chapter 1, the network topology and communication technology must also be carefully evaluated so as to mitigate the delay caused by communication as much as possible [68], [69].

Focusing on the control chain, a characterization of the operational delays of control systems using real-time digital simulator (RTDS) with hardware-in-the-loop (HIL) architecture is presented in [55]. A similar approach is followed in several scientific works including [177] (using OPAL-RT) and [178], [179] (using RTDS), but, in this literature, no measurement tool is provided specifically for the latency of the WAMPAC system and no metrological assessment is reported. In [180], the Electric Power Research Institute (EPRI) reports the results of experiments involving two ad-hoc modified PMUs, an Event and Time monitor and a Phasor Data Concentrator (PDC). Using different transport protocols, the delays of the event detection are computed with resolutions equal to the reporting interval.

5.2 Contributions of the overall WAMPAC latency

The considered WAMPAC architecture, shown in Fig. 5.1, consists of PMUs scattered throughout the territory, a WAMPAC core (referred to as WAMPAC

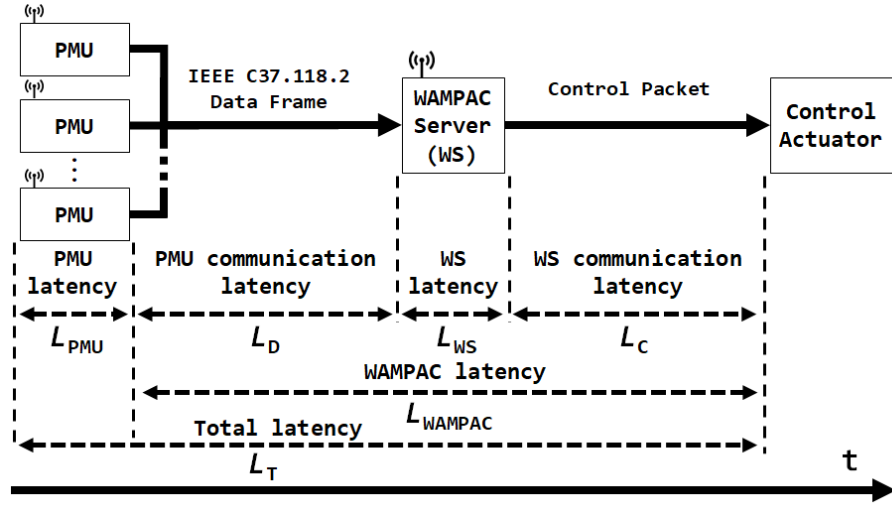


Figure 5.1: Diagram of the considered architecture of a generic WAMPAC system, composed of PMUs, WAMPAC Server and Control Actuator, with the associated latencies.

Server, WS from here on) and a module representing the control actuator. This is a common scheme and is used throughout this chapter for presentation purpose. Specific adaptations of such architecture can be also addressed with slight variations of the proposed approach.

As mentioned in Chapter 1, PMUs installed in the transmission network are scattered in the substations and can be configured according to the P and M compliance classes of the PMU Std [17]. This is a relevant scenario for WAMPAC applications and will be used as a typical situation in the rest of the chapter. P-class is usually recommended for WAMPAC applications since it requires fast measurement responses, while M-class is preferable in cases where higher measurement accuracy and greater rejection against noise are desired. For this reason, algorithms conforming to P-class usually require fewer fundamental cycles for measurement computation than those conforming to M-class, so they seem more promising for WAMPAC systems in order to reduce the latency of collected PMU data. The choice of the reporting rate (RR) of the PMUs is often critical for latency, because different algorithms and configurations can be used for different RRs.

For example, in the Italian WAMS there are 200 PMUs of different vendors installed in 400 kV and 220 kV substations. They are configured according to M-class and send the measurements every 20 ms, i.e., using a RR of 50 frames per second (fps), according to IEEE C37.118.2 protocol [36] through a private data network to Terna's National Data Center in Rome [37]. The IEEE C37.118.2

packets are typically received by an *ad hoc* PDC-enabled device (the WS in the considered architecture) that time-aligns the packets according to their timestamps and then, in the WAMPAC architecture, analyzes the incoming data and sends a control packet according to several possible implementations. For example, it may send a command only when the PMU measurements satisfy a certain WAMPAC policy. The packet sent by the WS can be of various types, such as a Routable-Generic Object Oriented Substation Event (R-GOOSE) message conforming to IEC 61850-90-5 [143], like in the WAMPAC system in [181], or an IEEE C37.118.2 packet as those sent by the PMUs.

The IEEE C37.118.2 Standard specifies the various packet types or frames for real-time communication between devices in a distributed measurement architecture. The standard outlines frames of the header, data, command, and configuration types. The focus here will be on the data frames, which contain the measurements collected from the PMUs, as well as command frames. Command frames are primarily used to initiate communication between PDCs and PMUs or to request information from the devices. In a pure monitoring context like WAMS, the role of the command frame is considered limited to the basic functionality defined in the IEEE C37.118.2 protocol. However, the standard defines seven standard commands (turn on transmission, turn off, send configuration frames, etc.), with some bits reserved for user designation. Wide-area applications that rely on synchronized measurements may thus use updated versions of the command frames for specific control applications.

An updated IEEE C37.118.2 command frame with a specific command for communication between WS and control actuator is considered in this chapter as an example of possible implementation. Such command frame allows a PDC to evolve into an active device for power system control. This message can be used in many applications as the activation signal for the actuator of the control (see Fig. 5.1). Otherwise, the content of the control packet can be transformed by adapters/gateways into various analog control signals if a WAMPAC system requires it (e.g., current signals in the 4-20 mA range as in [181]). The control actuator, based on the PMU measurements and the policies implemented by the WS, allows physical intervention in the system according to the required strategy.

The total latency L_T of the system, schematized in Fig. 5.1, is given by:

$$L_T = L_{PMU} + L_{WAMPAC} \quad (5.1)$$

with:

$$L_{WAMPAC} = L_D + L_{WS} + L_C \quad (5.2)$$

where:

- PMU reporting latency L_{PMU} in Eq. (5.1) is the time interval between the measurement time as indicated by the data timestamp and the instant when the data becomes available at the PMU output [17];
- WAMPAC latency L_{WAMPAC} is defined in Eq. (5.2) as the sum of the following terms:
 - PMU communication latency L_D is the time needed for the network to transmit the IEEE C37.118.2 data frame from the PMU to the WS;
 - WS latency L_{WS} is the time required for the WS to process data from the PMUs and send a control/command message;
 - WS communication latency L_C is the delay of the control/command message sent by the WS over the network.

The latency of the PMU is not taken into account in L_{WAMPAC} , as it depends mainly on the instrument, and the objective of latency characterization is twofold: find the total latency and find the latency introduced by WAMPAC system regardless the chosen commercial PMUs or their configuration. It is worth recalling that L_{PMU} depends on the performance class and RR considered, and its maximum value is recommended in [17]. For this reason and since L_{PMU} is usually measured in advance by manufacturer or laboratory characterization, here the focus is on L_{WAMPAC} . As it will be described in what follows, the designed characterization platform uses the IEEE C37.118.2 PMU packet output time (referred to as PMU sending time in the following) for the calculation of L_{WAMPAC} .

It is worth noticing that in a distributed PMU framework, i.e., in the presence of Stand-Alone Merging Units (SAMUs) or instrument transformers with digital output as seen in Chapter 4, the contributions given by the latency of the digitizer and the communication latency of the packets carrying the sampled values sent to the PMU [182] affect L_{PMU} . Therefore, these delays have to be characterized to understand their weight.

L_D and L_C also include the latency caused by devices, like routers, in the communication network. Moreover, L_D may be different between PMUs because they are located at different nodes in the network.

L_{WS} can in turn be divided into two contributions, according to needed WS functionality operation [18]:

- the “WS wait time” starts when the first PMU message with a given timestamp arrives and ends when the last message is received or the configured maximum wait time expires, after which the partial dataset is handled.

The difference between the end of such wait time and the arrival time of the last on-time PMU packet contributes to L_{WS} ;

- the “WS processing time”, usually quite smaller than the first, represents the time for the WS to process and complete the output stream. In WAMPAC applications, this interval includes WAMPAC policy evaluation.

As a general consideration, if the architecture is more complex than that in Fig. 5.1, e.g., if several PDCs are present in the hierarchy, the latency of each device and the communication latency of their sent packets is also considered in L_{WAMPAC} . Furthermore, if the message sent by the WS needs to be transformed into an analog signal, the latency given by the transducer device, which is usually negligible with respect to other contributions, should also be considered.

5.3 Proposed latency characterization system

In order to evaluate the latency of a generic WAMPAC system, as presented in [183], a “WAMPAC Characterizer” (also briefly “Characterizer” from here on) has been designed and implemented to emulate the behavior of different PMUs. Indeed, the Characterizer allows sending several packets compliant with the IEEE C37.118.2 Standard as if they were originated from a set of PMUs. In each PMU stream, packets are sent every T_{RR} seconds, i.e., according to the configured PMU RR. On the other hand, the Characterizer allows receiving control packets from the WAMPAC system, thus closing the control chain and allowing latency evaluation. As mentioned above, control packets are here assumed, for the sake of presentation clarity, to be formatted according to IEEE C37.118.2, which is a realistic choice in many applications. This choice goes in the direction of industrial best practice [59] and the typical scenario is that of TSO private or virtual private communication networks. The Characterizer matches the PMU communication capabilities and, in this regard, if additional security reinforcement approaches (e.g., security gateways) are used for PMU data, they can be applied straightforwardly also to the Characterizer output. It is foreseeable that, in future scenarios, when security protocols may be applied directly to PMU streaming [184], dedicated solutions can be used to extend the Characterizer capabilities in this perspective.

The WAMPAC Characterizer, which will be described in detail in Section 5.3.1, is an instrument devoted to WAMPAC latency measurement and thus needs in turn to be characterized to assess its performance before its application on the field. For this reason, a device called “WAMPAC Emulator” (or

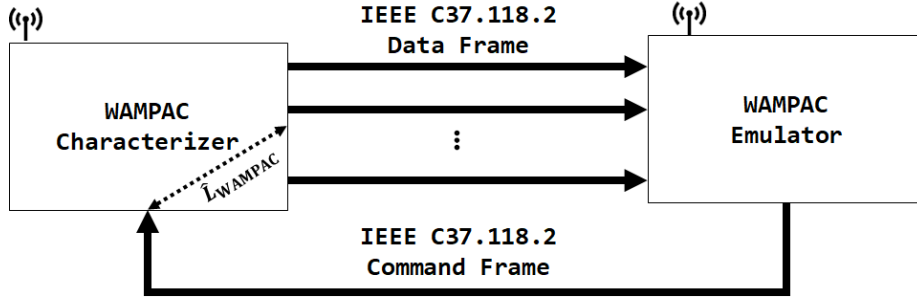


Figure 5.2: Schematic diagram of the laboratory characterization architecture.

“Emulator”) was also designed and implemented to physically emulate the behavior of a generic WAMPAC architecture with minimum additional latency.

Both the Characterizer and the Emulator allow configuring the number of PMUs, their RR and the PMU data frame size so that different scenarios of WAMPAC systems can be tested.

Fig. 5.2 shows the setup used for the laboratory characterization tests, in which there is a direct connection between the Characterizer and the Emulator to measure the latency required by a WAMPAC chain that mimics a real architecture without the impact of communication latency due to network devices and external network traffic. Through characterization, which will be discussed in Section 5.4.2, an evaluation of the internal latency of the Characterizer can be established in order to provide an indication of the measurement error introduced by the instrument. Then, the Characterizer can be applied to real WAMPAC scenarios (see Section 5.4.3).

5.3.1 WAMPAC Characterizer

The WAMPAC Characterizer allows the emulation of a variable number of PMUs (during the test, this number was raised up to 10, but more PMUs can be emulated depending on the computational capability of the considered hardware platform) capable of generating IEEE C37.118.2 packets with different packet sizes. User Datagram Protocol (UDP) or TCP can be chosen for the transport layer protocol as common to commercial PMUs. The RR of PMUs is also user-modifiable, and a RR up to 200 fps can be configured. Once again, higher RRs can be also used depending on the computational power available. These features make the WAMPAC Characterizer more flexible for testing WAMPAC systems with different requirements compared to the solution in [181], which allowed emulation of only 2 PMUs with RR at 50 fps and TCP messages of fixed packet size.

To minimize the intrinsic latency of the Characterizer, that inevitably affects

the measured WAMPAC latency, the emulated PMU packets are generated at Coordinated Universal Time (UTC) time instants corresponding to the PMU reporting instants (multiples of T_{RR} and written as the packet timestamps). In this way, packet flows emulate the real behavior of PMUs without introducing PMU measurement latency ($L_{PMU} \approx 0$). Moreover, to increase the level of emulation of the PMUs, the Characterizer can implement a specific delay for the sent packets of each PMU. In this way, real cases in which the PMUs do not have the same communication or reporting latency can be simulated.

Whenever packets have to be sent, the sending time of each packet ($t_{DS,i}$, where i represents the index of the emulated PMU) is obtained and written into the sent packet as an analog field. It allows having also the sending time available at the side of the receiver and not only the PMU timestamp (reporting time). Clearly, this option can be deactivated when the WAMPAC does not support the interpretation of this additional information, or the exact packet format of the real system needs to be used. After that, $t_{DS,i}$ of each PMU is stored in the system.

As mentioned above, the Characterizer is programmed to receive IEEE C37.118.2 packets from the WAMPAC system in response to PMU packets. Whenever a control packet is received, the current time (t_{CR}) is stored, and the difference between this arrival time and the largest timestamp of the corresponding PMU packets is calculated, thus computing the following latency:

$$\hat{L}_{WAMPAC} = t_{CR} - \max(t_{DS,1}, t_{DS,2}, \dots, t_{DS,i}, \dots, t_{DS,N}) \quad (5.3)$$

where $\hat{}$ denotes hereinafter measured quantities and N is the total number of emulated PMUs.

In addition, the Characterizer also computes the received control packet latency (WS communication latency)

$$\hat{L}_C = t_{CR} - t_{CS} \quad (5.4)$$

defined as the difference between the WS packet arrival time t_{CR} and the timestamp of the command frame, t_{CS} . In this way, the Characterizer calculates the latency of the packets coming from the WS, estimating the communication latency and allowing to highlight its contribution to the WAMPAC latency.

Fig. 5.3 shows the main functionalities of the Characterizer using a block diagram. The proposed instrument is designed to rely on real-time (RT) hardware and software and on a platform equipped with Field Programmable Gate Array (FPGA) to guarantee determinism in performance. These features are important to address timing and synchronization requirements and achieve high measurement accuracy as will be discussed in what follows. Configuration opera-

tions are performed on the parameters that can be set, such as the number of PMUs N , RR, packet payload size, protocol type and input data file. The actual synchrophasor streams of the various emulated PMUs can be chosen by the user based on an auxiliary file of measured values (e.g., 3-phase synchrophasor, frequency and rate of change of frequency (ROCOF) values for a certain time interval, collected from the field or simulated). The measured values can be used cyclically to test WAMPAC response repeatedly during long tests. The Characterizer can also provide simple measurement streams generated internally.

The main functionalities are divided into two parallel RT cycles, disciplined by the internal clock with a frequency equal to 10 MHz. The top RT cycle allows the set of PMUs to be emulated in terms of data transmission and timestamping. Indeed, the first block ensures alignment with reporting time instants by calculating the wait time required for alignment according to UTC. Whenever a time reference is needed, as seen in Fig. 5.3, the program makes a software call to the FPGA, which provides UTC time. Indeed, the FPGA is synchronized using an external time source; in this paper, a Global Positioning System (GPS) receiver is considered.

Within a number N of parallel instances, the UTC time is obtained and, consequently, an IEEE C37.118.2 data frame packet is built and sent with the corresponding reporting timestamp and, as mentioned above, with the measured sending time $t_{DS,i}$. The parallel instances are efficiently distributed according to the number of processors available in the platform adopted for implementation. At the output of the parallelized loop, a stop condition is evaluated: for instance, if the count of sent measurements is less than a set number (which depends on the RR, the length of the test, and the number of stream iterations), the RT loop is re-executed otherwise the loop terminates.

The bottom RT loop allows the command or data frame packets of the WS to be received. The possibility of receiving data frame packets was also included to allow testing even WAMPAC systems based only on a PDC and thus on data frames. The first block is responsible for receiving and checking these packets. When a packet is received, the current time is calculated and saved as t_{CR} and immediately afterwards the packet is decoded, extracting useful data such as t_{CS} . After decoding, $\hat{L}_{WAMPAC}$ and $\hat{L}_C$ are computed and stored according to Eq. (5.3) and (5.4), respectively. Lastly, a stop condition is checked.

The proposed Characterizer has additional features: it can be configured to be also used as an instrument to measure L_T , thus including PMU latency by checking the arrival time of the control packets with respect to the PMU timestamp. In this perspective, it allows testing mixed scenarios where the emulated PMUs and commercial PMUs coexist. L_{PMU} can be set for each emulated PMU in order to test different monitoring configurations.

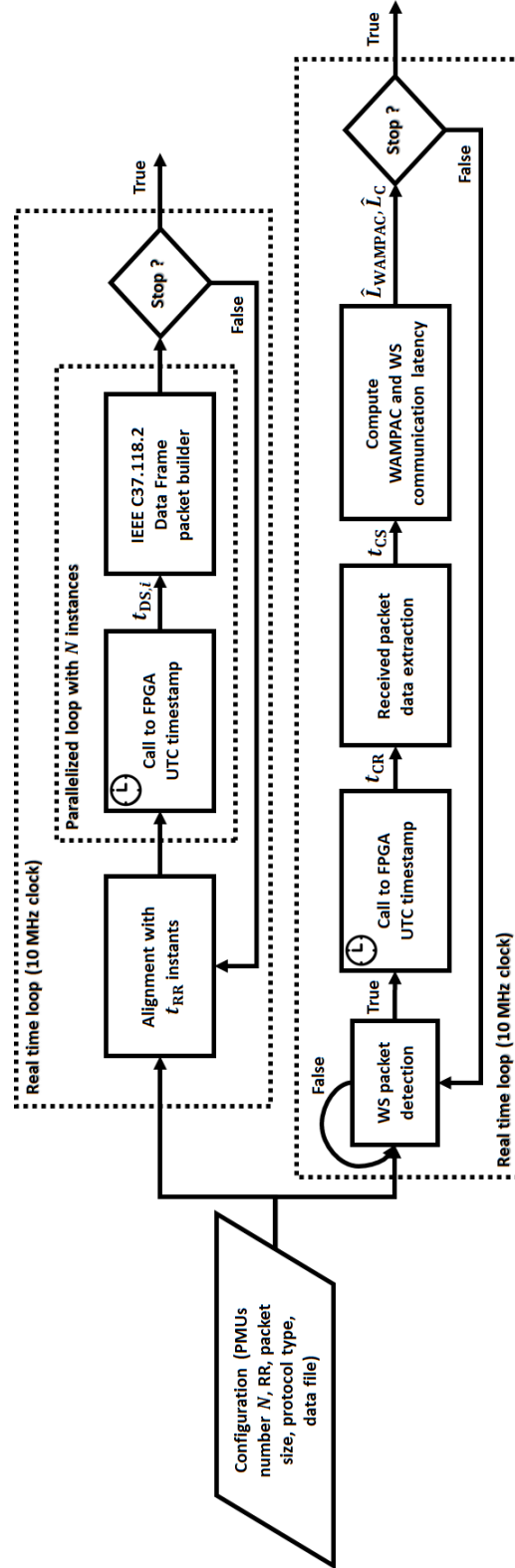


Figure 5.3: Block diagram of the proposed WAMPAC Characterizer.

The Characterizer was implemented in a National Instruments CompactRIO-9068 [158] (NI cRIO-9068), such as the one used in Chapters 3 and 4, therefore its performance will not be further described in this chapter. The NI cRIO-9068 was equipped also in this case with the NI 9467 module for GPS synchronization with a pulse per second (PPS) accuracy of ± 100 ns. In this paper, to reflect the behavior of digital WAMPAC systems, an IEEE C37.118.2 packet was chosen as the activation signal. Nonetheless, the modular and configurable hardware is ready to adapt the system to the needs of the specific WAMPAC system under test by changing the nature of the activation signal, for instance using a current signal in the 4-20 mA range (as used in [181]), or by changing the communication protocol, such as the IEC 61850-90-5 [143]. Indeed, at the end of Section 5.4.3, test results obtained when the Characterizer was also used to test the latency of a real Italian TSO WAMPAC prototype based precisely on 4-20 mA signals are reported and, for this reason and only for that test, the control signal to be received in the Characterizer is adapted to a current signal. In Section 5.4, this version of the proposed instrument will be referred to as the “4-20 mA WAMPAC Characterizer version”.

A specific version of the Characterizer software adapted to generic personal computers (PCs) was also implemented in order to provide another tool for preliminary latency assessment to TSOs and to evaluate the performance differences when using general purpose (GP) instead of RT hardware. This version differs from that used in the RT platform in terms of timestamping, since, in the absence of the FPGA, a software call to the computer time is used. The PC clock is synchronized via Network Time Protocol (NTP). For instance, in the tests of Section 5.4 below, the synchronization is provided by a Meinberg Lantime M1000 GPS receiver within the same Local Area Network (LAN) of the PC. Due to the non-RT behavior, it is not possible to have a highly accurate time and PC clock offset may drift up to 10 ms from the reference. However, the detected drift has a negligible impact on WAMPAC latency measurements when no strict synchronization is required by WAMPAC operation, and mainly affects communication latency measurements. For instance, if the Emulator is used, packet communication latencies to and from it are calculated based on timestamps from different timescales (internal computer clock and WS clock).

5.3.2 WAMPAC Emulator

The WAMPAC Emulator allows the emulation of fundamental WAMPAC functionalities to test the behavior of the Characterizer. Indeed, the Emulator implements the basic functions of the WS and can thus serve as a low-latency version of a real system. Therefore, the WAMPAC Emulator is capable of receiving the

IEEE C37.118.2 packets from PMUs (both real and emulated) and can run in real time a WAMPAC logic. Indeed, a policy must be chosen to decide when the activation signal has to be sent (in an IEEE C37.118.2 command frame packet in this chapter) for controlling the grid.

An example of WAMPAC policy based on phase-angle measurements, which is also adopted in the tests of Section 5.4, is the following rule:

$$\text{activation} = \begin{cases} \text{true} & \text{if } \max_{i=1,\dots,N} |\varphi_i - \varphi_{i_{\text{ref}}}| > \delta_\varphi \\ \text{false} & \text{otherwise} \end{cases} \quad (5.5)$$

where the activation signal is generated and the corresponding command frame is sent if the phase-angle difference between the phase-angle measurement $\varphi_{i_{\text{ref}}}$ of a PMU i_{ref} , chosen as reference, and at least one of the measured phase angles φ_i included in the other PMU packets with the same timestamp is greater than a threshold δ_φ . On the contrary, i.e., under normal conditions, the command packet is not sent.

The simple phase-angle monitoring (PAM) application considered here is inspired by that used by the Italian TSO in [181]. Furthermore, this simple policy appears effective when some real cases, such as the incident that occurred at the Continental Europe Synchronous Area (CESA) on January 8th 2021, are considered [185]. In that situation, shortly before the failure, the system was already operating close to the point of angular instability, with voltage phase-angle differences close to 90° between Western and Eastern Europe [185]. This event led to the separation of CESA into two sub-areas and demonstrated once again the importance of monitoring phase-angle differences between voltage phasors measured with PMUs at different locations, as they can be considered an early index of power transmission network stress [42], and thus may help in taking automatic countermeasures through purposely designed WAMPAC systems.

The Emulator is also capable of calculating the IEEE C37.118.2 packet latency $\hat{L}_{D,i}$ for the incoming PMU streams defined as the difference between the arrival time of input packet $t_{DR,i}$ and the sending time $t_{DS,i}$ written into the same packet, i.e.,

$$\hat{L}_{D,i} = t_{DR,i} - t_{DS,i} \quad (5.6)$$

In this way, the Emulator can analyze the behavior of the PMUs emulated by the WAMPAC Characterizer in more detail, providing additional insight. The Emulator also estimates its own processing latency $\hat{L}_{WE}$ as:

$$\hat{L}_{WE} = t_{CS} - \max(t_{DR,1}, t_{DR,2}, \dots, t_{DR,i}, \dots, t_{DR,N}) \quad (5.7)$$

$\hat{L}_{WE}$, as the WS processing time defined in Section 5.2, represents the difference between the instant when the activation signal is sent (t_{CS}) and the arrival time

of the last PMU packet among those with the same timestamp that trigger the policy.

Analogously to that illustrated for the Characterizer, the block diagram of the Emulator implementation is shown in Fig. 5.4. Also in this case, configuration operations are carried out on the parameters that can be set. In the same way as the Characterizer, the Emulator uses an RT cycle, regulated by the internal clock at 10 MHz, containing an internal cycle with N parallel instances distributed appropriately to the available processors.

Within the parallelized loop, PMU packets are received. The arrival time $t_{DR,i}$ is saved and the data frame is decoded, extracting $t_{DS,i}$. After obtaining the required data for all the incoming PMUs, the WAMPAC policy is applied (voltage and current synchrophasors, frequency and ROCOF can be considered according to the adopted strategy). If the policy condition is verified, a timestamp is recorded and the command frame is built.

The output packet was assembled according to the IEEE C37.118.2 Standard [36], using the bits of the CMD field reserved for user designation (identified with bits 15-12 set to 0 and bits 11-8 not equal to 0), to obtain a specific command for WAMPAC application.

Therefore, in accordance with the WAMPAC policy summarized in Eq. (5.5), this functionality has been used to send the control packet to activate the control strategy.

After the packet is sent, the WAMPAC Emulator processing latency $\hat{L}_{WE}$ and the PMU communication latencies $\hat{L}_{D,1}, \hat{L}_{D,2}, \dots, \hat{L}_{D,i}, \dots, \hat{L}_{D,N}$ are calculated, according to Eq. (5.7) and (5.6), respectively. If the policy results in a false condition, only the PMUs communication latencies are computed. Lastly, the RT cycle is re-executed according to the stop condition.

For the implementation of the WAMPAC Emulator, another modular system was used, the NI cRIO-9039 [186], provided with NI Linux real-time Operating System (OS), Intel Atom quad-core processor with maximum frequency 1.91 GHz, and 2 GB RAM, and equipped with the NI 9467 module for GPS synchronization. As in the WAMPAC Characterizer, the synchronization allows timestamping the packets arrival and departure with respect to UTC and consequently to calculate the aforementioned latencies in a coherent way.

Fig. 5.5 shows the modular devices used for the test setup during laboratory characterization and highlights the involved connections.

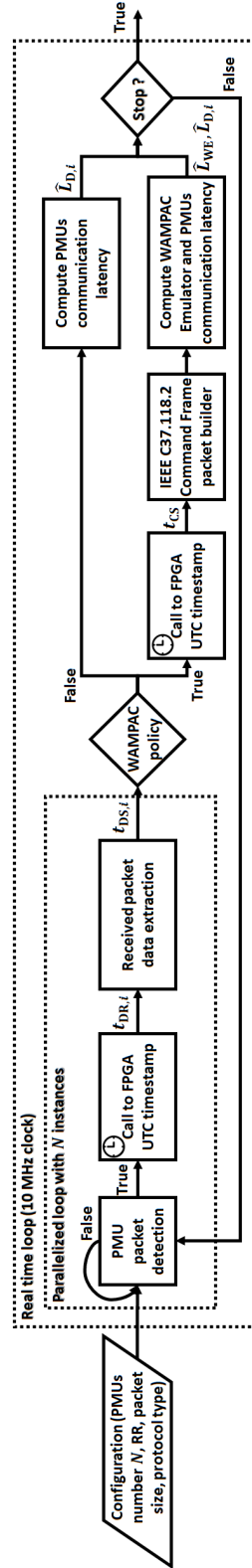


Figure 5.4: Block diagram of the WAMPAC Emulator.

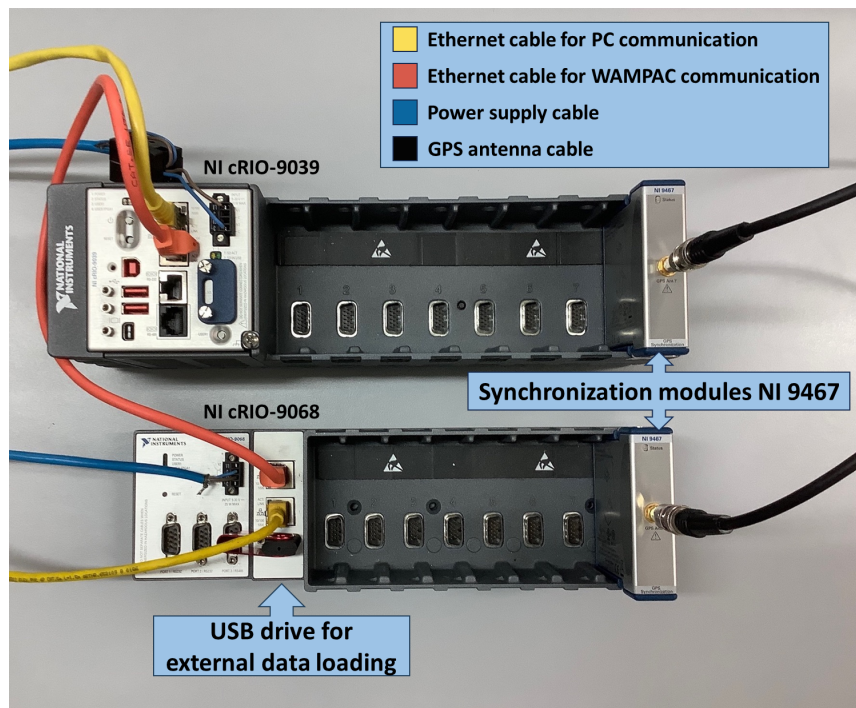


Figure 5.5: Picture of the used hardware and connections for the laboratory characterization.

5.4 Tests and results of WAMPAC latency measurements

In this section, two selected laboratory test scenarios and a test in a real WAMPAC prototype of the Italian TSO with the corresponding test setup are presented. Then, the most significant results are reported and discussed.

5.4.1 Test scenarios and setup

Laboratory evaluation of Characterizer latency

The first test scenario corresponds to the Characterizer latency laboratory assessment involving the Emulator introduced in Section 5.3 and, in particular, in Section 5.3.2. The test setup corresponds to Fig. 5.2. In the performed tests, PMU data correspond to synthetic synchrophasors generated directly by the WAMPAC Characterizer. Measured values were kept stable: for an emulated PMU assumed as reference, phase angle was constantly equal to 0° while, for the other emulated PMUs, it is kept at 110° , so that the phase-angle difference was always greater than $\delta_\varphi = 100^\circ$, which is considered as the event detection threshold. In this condition, the WAMPAC Emulator sends, for each set of packets from the emulated PMUs (each timestamp), a control packet. This allows characterizing the device in the worst stress condition, when it produces a latency measurement approximately every T_{RR} . This choice also permits a statistical assessment of latency measurement characteristics within short time intervals and is thus adopted from here on for the tests.

Experimental evaluation with an OpenPDC-based WAMPAC system

The second test scenario is an experimental latency measurement test for a WAMPAC system based on the OpenPDC tool by Grid Protection Alliance (GPA) [187]. This is an open-source software application intended as the core module in the WAMS and WAMPAC systems design. The OpenPDC is a data concentrator whose primary function is to time align and collect the synchrophasor data provided by the PMUs. It enables the management of data from PMUs in the field by also allowing outbound streams to be sent, concentrating data from the various incoming packets. Due to its capabilities, this software has been used in various scientific projects and real-field implementations [188], [189]. The test architecture is shown in Fig. 5.6(a). The OpenPDC is hosted in a workstation equipped with two Intel Xeon Gold 6128 processors, 256 GB RAM, and Windows 10 OS.

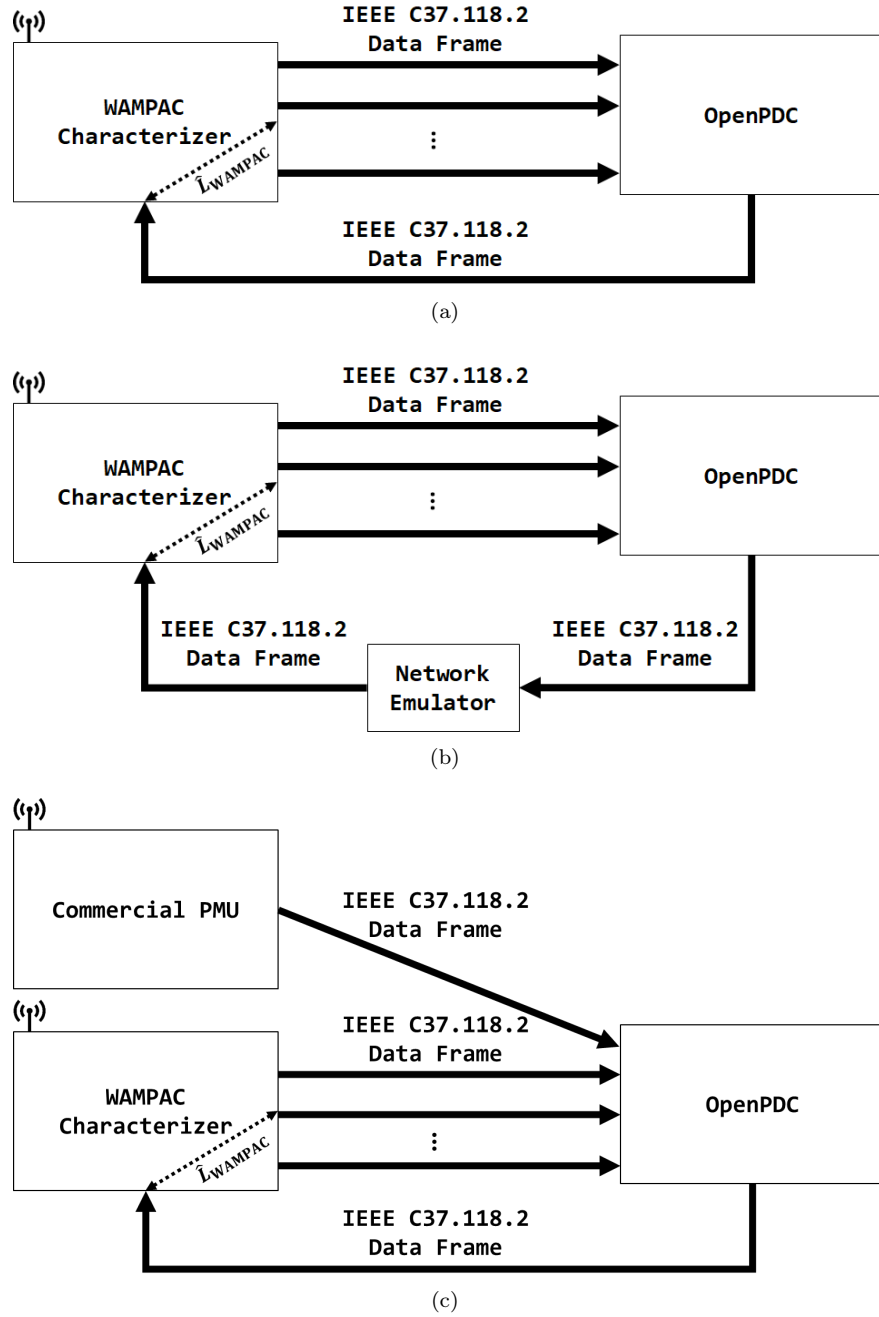


Figure 5.6: Schematic diagram of the experimental test architecture (a), integration of the network emulator (b), and of a commercial PMU (c).

To simulate various behaviors of a real network, a network emulator has been also used within the previously described architecture, thus modified as shown in Fig. 5.6(b). It is then possible to test applications in the presence of real network issues such as higher latencies, jitter and packet loss conditions. Network emulation with a similar approach has been used in several scientific works [47], [190]. The network emulator has been hosted in a workstation, with an Intel E8500 processor, 8 GB RAM and a Linux OS (Ubuntu 16.04 LTS), running the NetEm software [191]. The adopted workstation is equipped with several Ethernet network interfaces; two of them have been configured in transparent bridge mode, in order to influence the input and output data streams. As shown in Fig. 5.6(b), the network emulator has been connected between the WS and the Characterizer to simulate an additional latency to reach the actuator.

An additional test setup, shown in Fig. 5.6(c), has been used to test the same WAMPAC architecture in a mixed scenario, i.e., with the Characterizer and real PMUs connected to the same OpenPDC. In this case, the aim is to measure L_T , thus including the PMU reporting latency of a commercial PMU and the latency of PMUs emulated by the Characterizer.

Validation of a real Terna's WAMPAC prototype

Finally, the last test was carried out on a real WAMPAC prototype developed by the Italian TSO, Terna SpA, which needed to be validated from the latency point of view. Fig. 5.7 shows the main elements of this experimental characterization. In particular, this architecture is composed of the 4-20 mA WAMPAC Characterizer version introduced at the end of Section 5.3.1, a PDC located in Rome and a device called Smart Box (SB) located at Terna's laboratory in Cagliari. The last two devices will be treated as black boxes, i.e., they will be described only in terms of general functionalities. In this case, the WAMPAC Characterizer aims at emulating the behavior of just two PMUs in the same device. It sends 2 packets every 20 ms (one packet for PMU 1 and one for PMU 2) compliant with the IEEE C37.118.2 Standard, and receives current signals in the range 4-20 mA, thus closing the control chain.

The packets of the simulated PMU are received by the WAMPAC PDC, which typically sends every second a R-GOOSE message packet compliant with IEC 61850-90-5 [143] to the SB. The R-GOOSE message includes diverse information, depending on the incoming measurements and on the adopted WAMPAC policy. In particular, the WAMPAC PDC applies a configured policy (details are given in what follows) to the two incoming streams from WAMPAC Characterizer and generates the R-GOOSE output accordingly.

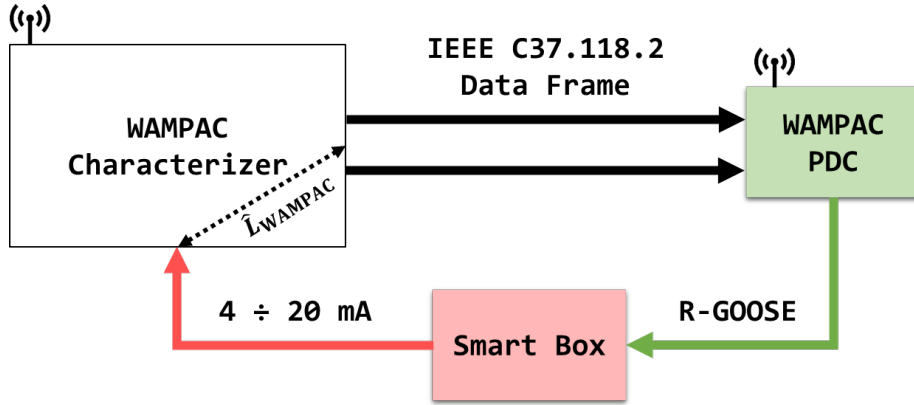


Figure 5.7: Diagram of the test architecture for the Italian TSO's WAMPAC prototype validated with the 4-20 mA WAMPAC Characterizer version.

The SB is in charge of receiving the R-GOOSE messages and acting as the generator of an analog command signal for WAMPAC application. In particular, a classic current loop signal in the range 4-20 mA is generated according to the content of R-GOOSE request. In the general configuration of Terna's WAMPAC, the 4-20 mA signal is considered the activation signal for the actuator of the control.

Again, a PAM application was considered with the same rule presented in Eq. (5.5) and with the same threshold value of $\delta_\varphi = 100^\circ$ as in the previous tests. The only difference is that the PDC changes a specific field in the R-GOOSE package where it changes the output of the SB from 4 to 20 mA when the threshold is exceeded. Thus, the PDC returns to the 4 mA value when it detects measurements within the threshold. Again, the policy is elementary because the idea is to focus on the temporal behavior of the architecture, but the characterization process can easily be customized for other types of applications.

For this reason, also in this test, the phase angle of the synchrophasors generated by PMU 1 was kept constant at 0° , while those of PMU 2 change with each UTC minute of occurrence, transitioning from a phase-angle value of 0° to a value of 110° to exceed the threshold. After the transition, the phase angle value is maintained for 10 s, after which the measured value returns to 0° . Exactly 500 values (50 fps for 10 s) with a phase value of 110° will then be sent from PMU 2 every minute. This transition for the phase and thus for the 4-20 mA signal was chosen in order to identify exactly when the PDC-SB chain comes into action. For this reason, compared to testing with the architectures in the previous laboratory scenarios, in this case a WAMPAC latency data is obtained every minute and therefore longer tests are preferable.

5.4.2 Metrological validation of the WAMPAC Characterizer

The first series of tests is devoted to the characterization of the metrological performance of the WAMPAC Characterizer in a controlled laboratory environment (see Section 5.4.1). The tests have been divided into short and long tests.

The Characterizer was initially verified with 15-minute (short-term) tests in which all configuration parameters were varied in order to check the instrument operation for every possible combination. Specifically, the following options have been used:

- 2 to 10 PMUs;
- RRs of 10, 25, 50, 100 and 200 fps, i.e., with a packet every 100, 40, 20, 10 and 5 ms. This corresponds to 9000, 22500, 45000, 90000, and 180000 measures for the 15-minute tests;
- 3 packet types with small (50 bytes), medium (94 bytes) and large (458 bytes) payload size. These different sizes mostly depend on the number of synchrophasors included in the data frame (3, 6 and 45, respectively);
- UDP or TCP transport layer protocol.

For the characterization tests, the architecture presented in Fig. 5.2 has been used to assess the Characterizer latency measurement accuracy. In particular, the results are summarized through the maximum value (indicated as ‘Max’ in the following tables), the mean value (μ) and the standard deviation (σ) of $\hat{L}_{\text{WAMPAC}}$ values provided by the Characterizer. Tables 5.1 and 5.2 present the results of the tests performed using the NI cRIO-9068, RR of 50 and 100 fps, and 2 or 10 PMUs.

In particular, Table 5.1 reports the results obtained when the UDP protocol is set, while Table 5.2 refers to TCP protocol. It is possible to observe the remarkable stability obtained with RT hardware: up to a RR of 100 fps, a maximum latency of about 1 ms was measured in the UDP case and about 1.3 ms in the TCP case. It is important to highlight that the obtained latency values include not only the effect of the Characterizer, but also the latency concerning the Emulator, thus guaranteeing that found maximum values are actually upper bounds for the latency introduced by the Characterizer.

It can be seen from the test results that μ remains almost constant as the number of emulated PMUs increases in the RT case. This, considering also that σ is practically constant too, means that the RT device, even if the number of operations to perform is higher, manages to keep its performance almost

Table 5.1: WAMPAC Latency Results for UDP Protocol and RT Hardware

Configuration		Reporting rate [fps]					
		50			100		
Number of PMUs	Payload size [B]	Latency index [ms]					
		Max	μ	σ	Max	μ	σ
2	50	0.90	0.64	0.03	0.93	0.64	0.03
	94	0.89	0.65	0.03	0.90	0.65	0.03
	458	0.96	0.70	0.03	1.02	0.69	0.03
10	50	0.87	0.64	0.03	0.95	0.63	0.03
	94	0.96	0.65	0.03	1.00	0.64	0.03
	458	0.98	0.70	0.03	1.05	0.70	0.03

Table 5.2: WAMPAC Latency Results for TCP Protocol and RT Hardware

Configuration		Reporting rate [fps]					
		50			100		
Number of PMUs	Payload size [B]	Latency index [ms]					
		Max	μ	σ	Max	μ	σ
2	50	1.03	0.74	0.03	1.04	0.73	0.04
	94	1.03	0.74	0.04	1.01	0.74	0.04
	458	1.08	0.79	0.04	1.10	0.79	0.04
10	50	1.24	0.74	0.05	1.28	0.73	0.05
	94	1.26	0.75	0.05	1.38	0.74	0.05
	458	1.31	0.80	0.05	1.30	0.80	0.05

unchanged in the various tests until time constraints can be respected. This is also confirmed by Fig. 5.8, which reports the histograms of $\hat{L}_{\text{WAMPAC}}$ measured when $\text{RR} = 100$ fps, small payload size, UDP protocol, and 2, 5 and 10 emulated PMUs are considered (red, green and blue bars, respectively). From the figure, it can be seen that the histograms are mostly overlapping.

On the other hand, with a RR of 200 fps, the system exhibits a state of heightened strain due to an increased number of concurrent requests. The load level of the CPU shows peaks at 100% occupancy and, for this reason, the device has some slowdowns, leading to higher maxima of measured latency in some tests. These correspond to few outliers that do not significantly affect μ and σ but reduce Characterizer's capability to accurately monitor a given system. Higher RRs are not manageable with the considered RT platform.

Tables 5.1 and 5.2 also show that, as the packet size increases, μ slightly increases, as expected. In addition, μ is generally lower using UDP protocol

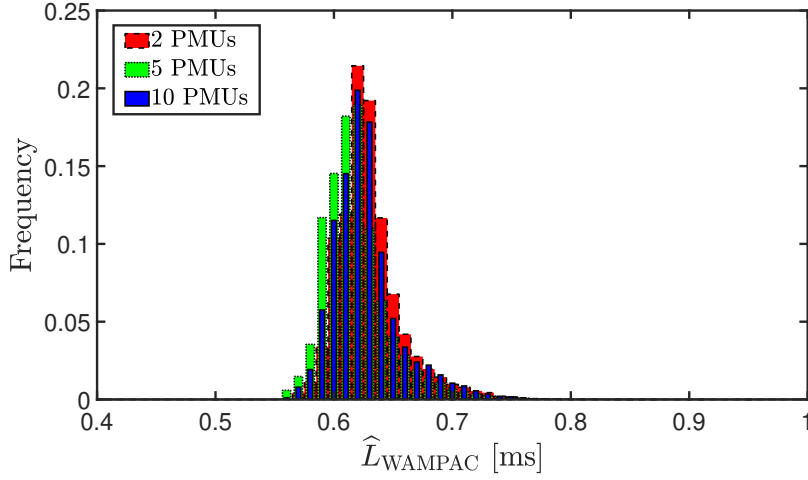


Figure 5.8: Histograms of the WAMPAC latency measured for 15 minutes, considering 2, 5 and 10 PMUs, RR at 100 fps and UDP small packet size.

than using TCP. This is because UDP is a lighter and faster protocol than TCP. Furthermore, the TCP packet usually has a larger header than UDP packet (in our case, the TCP header is 20 bytes while that of UDP packets is 8 bytes) considering the same number of synchrophasors transmitted.

Table 5.3 shows an extract of the results achieved with GP hardware with UDP protocol and the same test length (see Table 5.1 for comparison). For these tests, the Characterizer software was run in a workstation equipped with two Intel Xeon Gold 6128 processors, 256 GB RAM, and Windows 10 OS. To synchronize the workstation, NetTime software [192] with a 15-minute refresh interval was used. The results are much more variable and maximum values are quite pronounced due to frequent interrupts that can slow down code execution (the worst found result is about 90 ms for 7 PMU, small packet and RR = 200 fps). In this case, σ reflects the increased variability, whereas μ can be even lower in some tests than that obtained with RT hardware. Indeed, the used workstation provides more computing power than the employed RT system but no determinism is guaranteed. As for TCP, all these considerations still hold but, as in the RT case, with generally higher μ values than UDP (the results are not reported here for brevity).

To clarify the differences in the results obtained with the RT device and the GP device, Fig. 5.9 shows the trends of the measured WAMPAC latency in case of 2 emulated PMUs over 15 minutes. From the figure, the higher stability of the values obtained with RT hardware is evident, whereas GP latency measurements are characterized by the presence of outliers. Therefore, the characterization of a

Table 5.3: WAMPAC Latency Results for UDP Protocol and GP Hardware

Configuration		Reporting rate [fps]					
		50			100		
Number of PMUs	Payload size [B]	Latency index [ms]					
		Max	μ	σ	Max	μ	σ
2	50	30.10	0.55	0.18	11.52	0.47	0.07
	94	7.45	0.54	0.10	45.19	0.49	0.40
	458	8.64	0.50	0.09	4.49	0.48	0.06
10	50	24.52	0.81	0.15	9.17	0.73	0.09
	94	4.24	0.78	0.09	49.05	0.74	0.28
	458	26.64	0.80	0.18	35.50	0.76	0.19

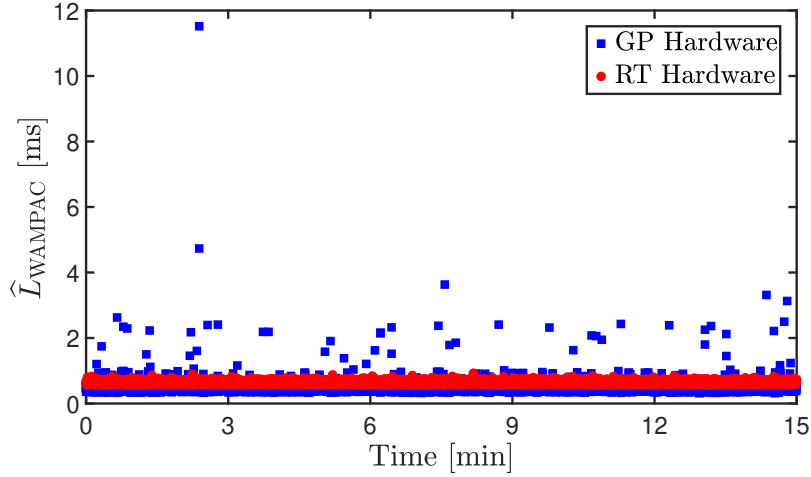


Figure 5.9: WAMPAC latency measured by RT and GP hardware, considering 2 PMUs, RR at 100 fps and UDP small packet size.

WAMPAC system using the RT system shows, as expected, stable performance over time. Nonetheless, the comparison reveals that the GP implementation, even if not highly accurate, can be considered a useful tool for preliminary compliance tests of WAMPAC functionalities and investigation of the system behavior.

To get further into the Characterizer assessment, the various contributions to $\hat{L}_{\text{WAMPAC}}$ are isolated and shown in Fig. 5.10, considering 10 emulated PMUs, 50 fps, UDP protocol and medium packet size. $\hat{L}_{\text{WAMPAC}}$ and $\hat{L}_C$ (latency of the control packet) measurements, obtained with the Characterizer, are reported as a function of time together with the $\hat{L}_{\text{WE}}$, measured internally by the Emulator, and the maximum $\hat{L}_{D,i}$ obtained by the Emulator among the PMUs. Table 5.4

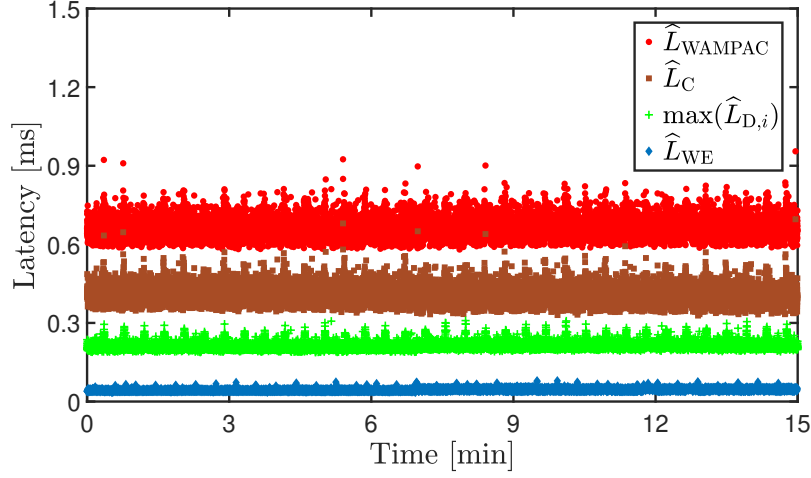


Figure 5.10: From the top to the bottom: WAMPAC latency (red dots), control packet latency (brown squares), max PMU communication latency (green crosses), and WAMPAC Emulator processing latency (blue diamonds), considering 10 PMUs, 50 fps and UDP medium packet size.

shows μ and σ of the individual latency components measured during the test. These results indicate that the latency introduced by the WAMPAC Emulator can be safely considered to be lower than 0.1 ms. Obviously, this latency contribution is related to the complexity of the policy and the number of PMUs, but, thanks to the RT performance, it is still an order of magnitude below other latency contributions.

Table 5.4: Latency Components for 10 PMUs, 50 fps, Medium Packet Size, and UDP Protocol

Latency Component	Latency Index [ms]	
	μ	σ
$\hat{L}_{WAMPAC}$	0.65	0.03
$\hat{L}_C$	0.40	0.03
$\hat{L}_{WE}$	0.04	< 0.01
$\max(\hat{L}_{D,i})$	0.20	0.01

To verify the functionality in the presence of a network issue, an extra communication delay in a test over 15 minutes has also been added. The above test configuration has been used and the data stream of the 10th PMU is delayed by 1 ms starting from the middle of the test duration and leaving the other PMU streams unaffected. The test shows that PMU communication latency (and consequently also WAMPAC latency) has an instantaneous increment equal to

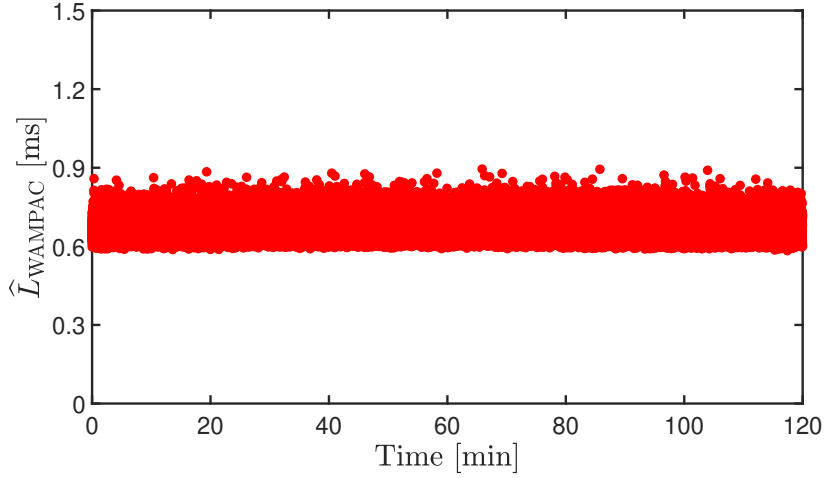


Figure 5.11: WAMPAC latency measured for 2 hours, considering 2 PMUs, RR at 50 fps and UDP medium packet size.

the extra delay, as expected.

The second set of characterization tests corresponds to long-duration tests, which have been carried out using the RT hardware. The Characterizer configuration is composed of 2 emulated PMUs, 50 fps, and medium packet size. In order to verify the stability and accuracy of the implemented system in the long run, 2-hour tests are considered with both UDP and TCP protocols. Fig. 5.11 reports the latency measured by the Characterizer for the UDP case and reveals that it remains stable even during long tests. Almost the same results as in Table 5.1, with values always below 0.90 ms (1.03 ms for TCP), were achieved: in particular, $\mu = 0.66$ ms and $\sigma = 0.03$ ms for the UDP case, while $\mu = 0.75$ ms and $\sigma = 0.03$ ms for the TCP case. The TCP case can also be compared with [181], leading to an accuracy improvement of one order of magnitude, from about 10 ms to 1 ms. Furthermore, these results can be used to interpret the accuracy of the measurements performed in the experimental tests of Section 5.4.3.

Afterwards, tests in the presence of network delay have been performed. A preliminary test was conducted to verify that the network emulator does not introduce significant extra delays (in addition to those set) into the measurement and control chain. For this reason, the network emulator has been configured to provide no extra delay. The test was carried out with both UDP and TCP protocols, but for the sake of brevity, only the UDP case is shown in Fig. 5.12 (results obtained with TCP are similar). In this test, $\mu = 0.70$ ms and $\sigma = 0.04$ ms have been obtained. As expected, both the mean and standard deviation

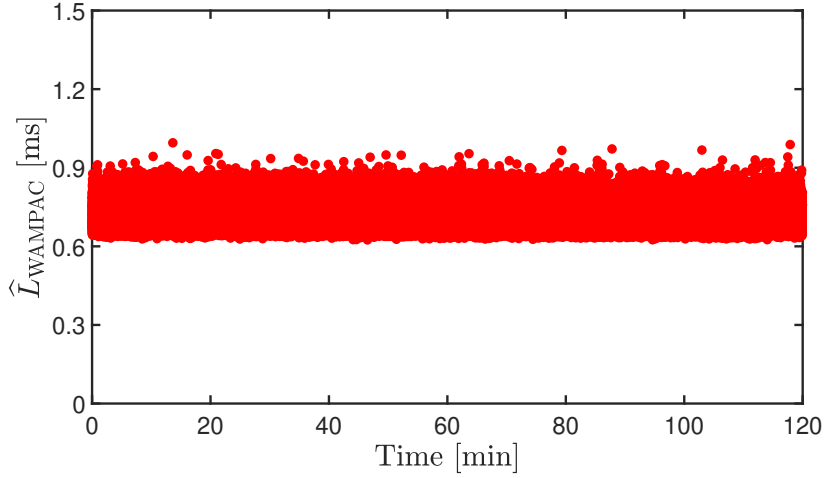


Figure 5.12: WAMPAC latency measured for 2 hours through the integration of the network emulator without extra delay, considering 2 PMUs, RR at 50 fps, and UDP medium packet size.

are slightly larger than in previous tests without the network emulator, but the additional latency is negligible (the network emulator introduces an average latency of approximately 0.04 ms).

Finally, two additional characterization tests were carried out with UDP and TCP protocols by setting a significant delay in the network emulator in order to emulate the communication delay between the WS and the WAMPAC actuator. Considering that the maximum delay of the round-trip time in a real test case reported in [181] was 36 ms, 18 ms has been chosen as a representative one-way network delay, thus emulating the latency of a real communication network. For the UDP case, $\mu = 18.70$ ms and $\sigma = 0.03$ ms have been achieved. As expected, the obtained average value is shifted by 18 ms compared to the previous ones.

In all the short- and long-duration tests, no packet loss occurred, which is a crucial aspect to consider for employing the instrument in the field.

5.4.3 Experimental tests on real WAMPAC systems

In this Section, experimental tests conducted on a WS based on OpenPDC and on a real WAMPAC prototype of the Italian TSO based on 4-20 mA actuator technology, both described in Section 5.4.1, are presented.

Latency characterization of OpenPDC-based WAMPAC system

First, the architecture shown in Fig. 5.6(a) has been used and 2-hour tests have been carried out with both UDP and TCP protocols. Fig. 5.13 reports the

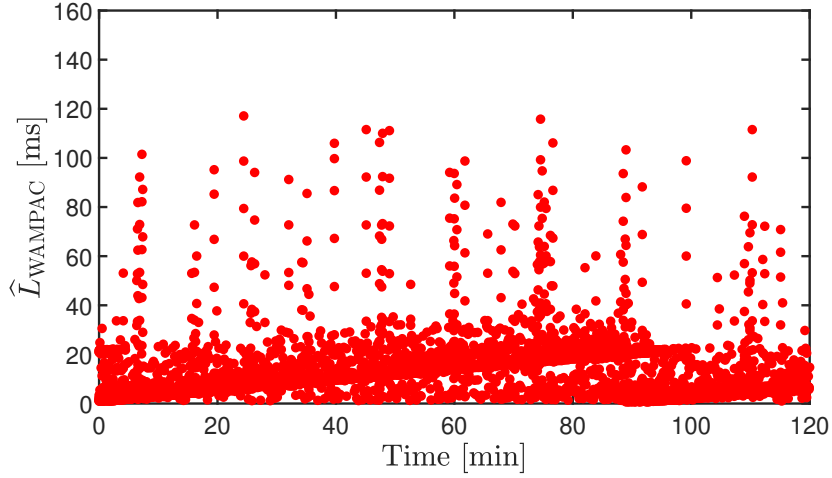


Figure 5.13: WAMPAC latency measured for 2 hours through the connection to OpenPDC, considering 2 PMUs, RR at 50 fps and UDP medium packet size.

results for the UDP case and a substantially similar behavior has been obtained with TCP. A linear trend with a positive slope can be observed interleaved with higher latency bursts. The linear trend lasts about 100 min and yields a variation of about 20 ms. This behavior is thus representative of OpenPDC operation. L_{WAMPAC} goes up to 120 ms in the UDP case and 140 ms in the TCP case for a few instants, while progressively smaller delays follow such spikes. In these tests, an average of 9.47 ms and a standard deviation of 6.16 ms were obtained for the UDP case, while $\mu = 12.19$ ms and $\sigma = 5.76$ ms were found in the TCP case. These results are intrinsically tied to the workstation and OpenPDC application configuration and characteristics. They also depend on the concurrent processes running on the workstation and can be partially improved by increasing the OpenPDC process priority. Several tests have been performed in this regard, highlighting the role of a non-real-time system in the latency trends. However, an analysis of the application configuration and performance is beyond the scope of this thesis.

Moreover, two tests were carried out with the architecture in Fig. 5.6(b), introducing an additional delay of 18 ms through the NetEm like in Section 5.4.2. The results are shown in Fig. 5.14 for the TCP case, leading to $\mu = 29.38$ ms and $\sigma = 5.33$ ms. The latency values appear substantially shifted by the added delay with respect to the prior results, but the pattern is the same as in the previous test.

A further test was carried out using a commercial PMU and two PMUs emulated by the WAMPAC Characterizer connected to the OpenPDC, as shown

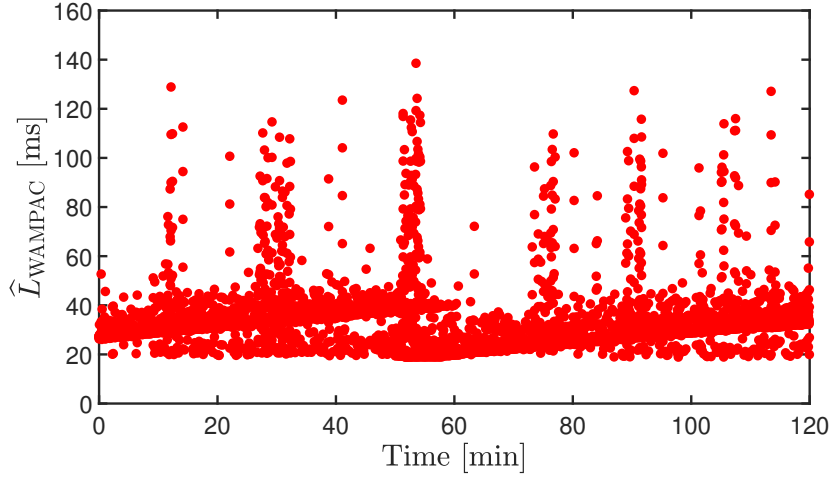


Figure 5.14: WAMPAC latency measured for 2 hours through the connection to OpenPDC and network emulator with an additional delay of 18 ms, considering 2 PMUs, RR at 50 fps, and TCP medium packet size.

in Fig. 5.6(c). The chosen commercial PMU uses an M-class algorithm for synchrophasor estimation with RR = 50 fps and TCP packets of 92 bytes. Such PMU was characterized beforehand in terms of reporting latency, finding an average $\mu_{\text{PMU}} = 132.43 \text{ ms}$ and a standard deviation of 0.08 ms, for about 30 minutes, since the PMU Std [17] requires at least 20 minutes of observation. Therefore, to make the scenario more realistic, the same latency was configured for the PMUs emulated by the Characterizer. The test results (not reported here for the sake of brevity) confirm the conclusions of previous tests (e.g., that in Fig. 5.13) while highlighting the Characterizer flexibility. Indeed, the average of $\hat{L}_T$ is almost shifted by μ_{PMU} with respect to $\hat{L}_{\text{WAMPAC}}$ of the above test, as expected, and the standard deviation is almost the same. Latency peaks are more erratic and depend on the specific test configuration, but the influence of L_{PMU} on the measurements is still evident.

Latency validation of a real Terna's WAMPAC prototype

In this section, tests carried out using the 4-20 mA WAMPAC Characterizer version proposed in [181] to validate Terna's WAMPAC prototype will be discussed.

At the implementation level, this involves only the modification of the “WS packet detection” block in Fig. 5.3, which is replaced by a current acquisition via the NI 9203 module (with a maximum sampling rate of 200 kS/s). In that case, the timestamp is saved whenever a transition from low level (4 mA) to

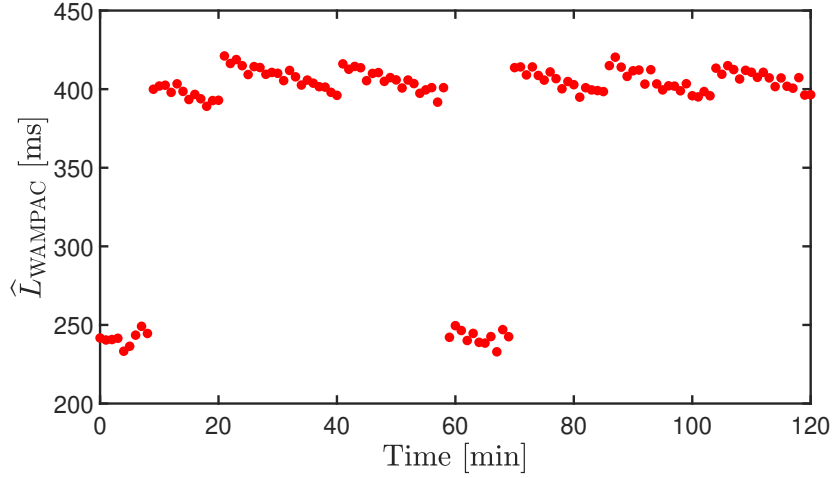


Figure 5.15: WAMPAC latency of the Terna’s WAMPAC prototype based on 4-20 mA actuator technology measured using the 4-20 mA WAMPAC Characterizer version for 2 hours, considering 2 PMUs, RR at 50 fps and TCP medium packet size.

high level (20 mA) is detected. This transition represents the activation of the policy in some WAMPAC systems, as in the one tested here. In addition, since there are no more packets to decode, no data are extracted from the received signal and the communication latency $\hat{L}_C$ of the packets cannot be estimated. It is important to highlight that this change has no detrimental impact on the latency of the WAMPAC Characterizer itself and by dealing with nondeterministic Ethernet communication at the receiver side the latency introduced by the device even decreases.

Passing to the validation of Terna’s WAMPAC prototype, a preliminary measurement of the round-trip time via ping command between Cagliari and Rome was made, in order to obtain an indicative information of the communication latency. Based on 100 packets, a minimum of 33 ms and a maximum of 36 ms were found. After that, using the architecture of Fig. 5.7, a 2-hour test was then carried out to measure the WAMPAC latency and a graph of the results is shown in Fig. 5.15.

As it can be seen from the trend, two “modes of operation” appear, with two different latency plateaus, which are characteristic of the specific WAMPAC PDC configuration set by the TSO. An average of 379 ms and a standard deviation of 60 ms were obtained.

In a second test, lasted ten hours, the antenna was disconnected and the Characterizer undergoes a gradual loss of accuracy of clock synchronization, which, however, should not affect the total latency measurements. This condi-

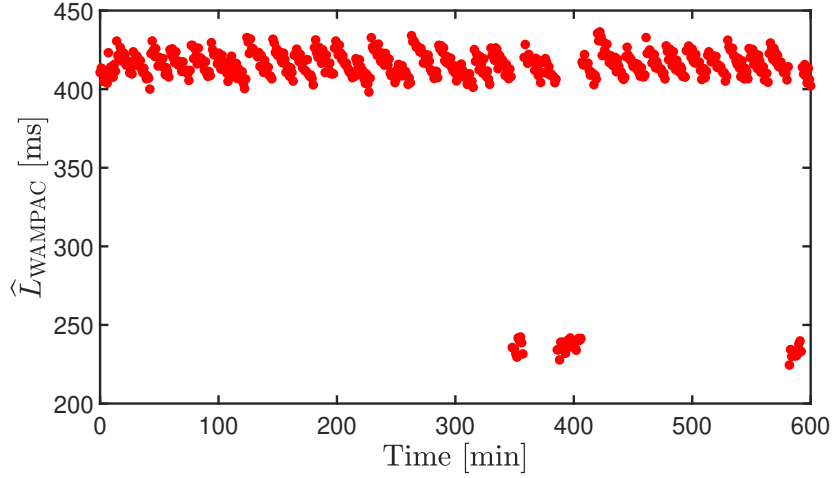


Figure 5.16: WAMPAC latency of the Terna’s WAMPAC prototype based on 4-20 mA actuator technology measured using the 4-20 mA WAMPAC Characterizer version for 10 hours with no time synchronization, considering 2 PMUs, RR at 50 fps and TCP medium packet size.

tion was tested to understand if absolute timing is critical for the WAMPAC system. The results of the WAMPAC latency for the 4-20 mA transition are shown in Fig. 5.16. The trend represented in Fig. 5.16 shows, similarly to the previous test with time synchronization, two states in which the overall latency settles alternatively around 240 ms and 410 ms. For this test, an average of 404 ms and a standard deviation of 46 ms were obtained.

It is important to point out that, the tests presented here using the architecture in Fig. 5.7 are dated back to the 4-20 mA WAMPAC Characterizer version, i.e., the first presented version of the instrument with an average total latency of 5.02 ms, a standard deviation of 0.52 ms, and an upper limit for introduced latency ≤ 10 ms. Thus, these results must be taken into account when evaluating the accuracy of the measurements made in the field. Nevertheless, the accuracy of this first version of the WAMPAC Characterizer is more than an order of magnitude lower than the latency measurements found, so it is adequate for this test.

Furthermore, for both tests, the latency seems to follow a piecewise linear trend, a sort of sawtooth wave, that was not found in the characterization of the implemented Characterizer and does not depend on its synchronization (Fig. 5.16 compared with Fig. 5.15). Such trend can be thus considered representative of Terna’s WAMPAC prototype.

In conclusion, in all the evaluated tests, the proposed measurement platform proved to be effective, with an accuracy well-suited for testing WAMPAC ap-

plications in real operating scenarios. This measurement instrument can then help TSOs during the prototyping and implementation phases, so that configuration and testing can rely on an accurately measured WAMPAC latency. This is a key factor, because, when WAMPAC latency is not assessed properly, automated control may not work and, in any case, cannot be optimized. The results also show that the GP version does not allow the accuracy required to verify real WAMPAC constraints, but it can help in preliminary tests. Finally, the last tests on the real prototype of WAMPAC system at Terna's highlight that a WAMPAC system should be characterized in terms of latency for long periods and under different conditions, because such a complex and interconnected system might reveal peculiar interactions among modules and unexpected behaviors that need to be well described before moving to the operation stage.

Conclusions

The ongoing evolution of the electric grid from traditional, centralized architectures to more decentralized, intelligent, and dynamic networks introduces significant challenges in terms of stability, control, and reliability. This transformation is largely driven by the increasing penetration of distributed generation, Renewable Energy Sources (RESs), and the electrification of all sectors of the economy, for instance in the transportation sector with the introduction of Electric Vehicles (EVs), which are adding complexity to the electric power system. The promising Wide Area Monitoring, Protection, and Control (WAMPAC) systems have therefore become indispensable tools for providing real-time visibility and adaptive control over these complex grids.

In this context, this thesis contributes to the design, development, enhancement, and validation of key measurement tools and methodologies of WAMPAC systems for modern power grids, focusing on synchrophasor estimation methodologies for Phasor Measurement Unit (PMU) valid in presence of dynamic and transient conditions, advanced White Rabbit (WR) synchronization technology, synchronization quality assessment in digital substations, and on the characterization of the overall WAMPAC latency.

In particular, Chapter 2 has introduced a new approach to synchronized phasor, frequency and rate of change of frequency (ROCOF) estimation that enables high measurement accuracy also during abrupt variations. It combines the advantages of Taylor-Fourier Multifrequency (TFM) models in terms of dynamic tracking and disturbance rejection, with the fast response to step events brought by asymmetric Taylor-Fourier expansions recombination. Simulation results have shown that the proposed approach opens many possibilities to define algorithms capable of dealing with abrupt, unexpected transitions. In particular, its variants can lead to the design of P-class compliant methods or M-class solutions, with remarkable performance in the presence of step changes. Zero or quasi-zero response time for synchrophasor, frequency and ROCOF can be even obtained under realistic noise conditions and for non-instantaneous changes of

the monitored quantity. This makes the proposal very promising to reduce the number of measurements labeled as “bad data” and thus discarded in real-world applications, and extremely useful for WAMPAC applications in power systems. Furthermore, as future developments, the results of the method can be further improved, especially from the point of view of noise robustness, even trying to further reduce the measurement window to limit PMU and thus WAMPAC latency. This idea is already in development and will lead to further innovation in the field of synchrophasor estimation.

Chapter 3 has discussed the contribution of time synchronization uncertainty achieved through WR technology in time-sensitive power system monitoring applications, like harmonic synchrophasor measurement and fault localization through traveling waves, and has proposed a possible architecture and easy-to-use hardware framework to achieve synchronization in this context. Real-time hardware that guarantees determinism and high-speed communication has been used for the implementation. To verify the achievable performance, architectures characterized by different synchronization accuracy specifications were compared. The results highlight the great stability and high accuracy achievable through the WR protocol, estimating a standard deviation of the offset from the time reference of about 10.5 ns, which may be further improved through higher performance hardware. In the first application scenario, the achieved synchronization accuracy determines a harmonic synchrophasor phase error contribution below 0.05 crad for harmonic orders up to the 50th. The second investigated application is the fault location through traveling waves. It has been estimated that the localization standard uncertainty for WR architecture emerges to be about 2 m, namely about an order of magnitude lower with respect to the conventional Global Positioning System (GPS) case. These results can be further improved by leveraging higher-performance hardware that can further exploit the potential of the sub-nanosecond accuracy achievable with WR technology.

Chapter 4 has considered the measurement chain of a PMU that receives as input the Sampled Value (SV) data stream output by a Stand-Alone Merging Unit (SAMU). Both the PMU and the SAMU are equipped with independent GPS antennas that guarantee a stable lock to Coordinated Universal Time (UTC). If the SAMU loses its lock to UTC, a strict application of the standards’ requirements would imply the entire measurement chain to be not synchronized, considering also the PMU measurements invalid until the full restoration of the SAMU synchronization. Such a process could take up to some tens of minutes, with severe consequences on the operation continuity of several PMU-based monitoring and control services (e.g. fault detection or state estimation). How-

ever, a similar approach may prove to be excessively conservative, since it does not take into account neither the remarkable stability of the internal clocks of modern SAMUs and PMUs (even when not disciplined by external sources) nor the potential benefits of such a redundant synchronization configuration.

In order to address this challenge, the thesis has proposed an alternative approach for the real-time update of the PMU Time Quality flag. Based on the assumption that the PMU internal time-base is still locked to a stable and traceable reference, it is possible to timestamp the incoming SV packets and evaluate their statistical distribution with respect to UTC time. This analysis is performed in real-time and allows for retrieving useful information about the SAMU synchronization status and update accordingly the PMU Time Quality flag. The performance of the proposed method was characterized in a plausible implementation of the full measurement chain by means of commercially available industrial controllers. The accuracy of the timestamping functionality was evaluated in the presence of both stationary and oscillatory publishing periods. Based on these tests, the PMU has been proven capable of timestamping the SV packets with a resolution of $1\text{ }\mu\text{s}$ and of tracking time-varying trends with oscillations up to few hertz. Then, a typical behavior of a SAMU losing its lock to the external time reference was reproduced and the proposed method allowed to re-assess the PMU Time Quality within $10\text{ }\mu\text{s}$ in few hundreds of seconds, i.e. several minutes before the SAMU is actually back to complete and stable synchronization. Even if this is neither an accurate nor a traceable measurement method, with a sufficient confidence interval it is possible to restart the PMU operation early. The overall accuracy would be sub-optimal, but anyway preferable than a complete absence of information.

Finally, Chapter 5 has proposed a WAMPAC Characterizer that is flexible and adaptable to the different types of WAMPAC systems. Its accuracy allows testing different WAMPAC scenarios, with several PMUs and with various reporting rates (RRs), packet configurations, and communication network delays. While evaluating the WAMPAC system under test and its promptness, the proposed platform does not introduce any significant uncertainty-related decision risks. The proposed measurement platform can be indeed employed for both short- and long-duration tests also when WAMPAC systems rely on UTC synchronization for their critical tasks and is thus a promising tool for transmission system operators (TSOs) at design, prototyping and pre-production stage. Indeed, the WAMPAC Characterizer was even employed for the validation of a real WAMPAC prototype of the Italian TSO, Terna, and has allowed to adequately characterize this system from the latency perspective.

All these advancements pave the way for more accurate and effective power grid monitoring and control solutions, addressing the evolving challenges of modern power systems. The insights and methodologies presented here will be valuable tools for future developments in WAMPAC technology, supporting the transition towards a more reliable and efficient power system infrastructure.

A recurring theme throughout this thesis has been the importance of minimizing measurement downtime and ensuring the validity of measurement data, even under challenging operational conditions. Whether through advanced algorithms for synchrophasor estimation, innovative approaches to synchronization quality assessment, or latency characterization systems, the proposed methods aim to maximize the availability and reliability of WAMPAC system outputs. This emphasis on providing valid or near-valid measurements, even in the presence of synchronization losses or abrupt transients, represents a guiding principle of the research. Reducing dead times and enabling more resilient measurement processes are critical not only for the operational continuity of modern power systems but also for supporting their ability to adapt and respond to rapidly changing conditions. These principles provide a unifying perspective for the contributions presented and underline the importance of reliable, continuous monitoring in the face of increasing grid complexity and variability.

Looking ahead, the future of WAMPAC systems will likely incorporate tighter integration with machine learning and artificial intelligence algorithms, enabling predictive diagnostics, self-healing capabilities, and improved resilience to cyber-physical threats. Further advancements in synchronization technologies, such as quantum-based time standards, could drive WAMPAC measurements closer to ideal accuracy, minimizing instrument timing errors. Additionally, as power grids become more decentralized, WAMPAC systems will need to manage higher levels of data processing and real-time decision-making at the edge.

In parallel, synchrophasor estimation methods will need to become even more accurate and robust under rapidly changing dynamics and frequent transients, which are increasingly present in modern power systems. These algorithmic improvements will be crucial to maintain adequate measurement accuracy and minimum response time during the complex and variable conditions characteristic of future grids. Together, these advancements will ensure that WAMPAC systems remain at the forefront of grid innovation, prepared to address the evolving challenges of tomorrow's energy landscape and support a sustainable and resilient power grid.

Appendix

Allan Deviation

Allan deviation (ADEV) is a statistical measure used to evaluate the stability of frequency signals over different time intervals. It is commonly applied in the field of time and frequency metrology to characterize the short- and long-term stability of oscillators, clocks and synchronization systems [159]. Unlike the standard deviation, which can diverge due to long-term drifts in the data, the Allan deviation is specifically designed to measure stability over specific time intervals, making it a valuable tool for analyzing systems where stability over different observation periods is essential [193].

In synchronization applications, Allan deviation helps to quantify the phase noise and frequency fluctuations of synchronization technologies, thus enabling the analysis of noise processes, such as white noise, flicker noise and random noise, which can introduce timing errors into synchronization signals compromising their performance [194].

To understand the Allan deviation, it is important to first define the fractional frequency deviation. The fractional frequency deviation $\bar{y}_i$ is calculated over a specific averaging interval τ , which represents the duration over which the frequency is averaged. It is defined as the difference between the signal's average frequency over τ and the nominal frequency f_0 , normalized to f_0 . Note that the bar over the y symbol is used here to denote the average. Mathematically, if $f(t)$ is the instantaneous frequency of the signal, then $\bar{y}_i$ over an interval starting at time t_i is given by:

$$\bar{y}_i = \frac{\langle f(t) \rangle_{t_i}^{t_i+\tau} - f_0}{f_0} \quad (\text{A.1})$$

where $\langle f(t) \rangle_{t_i}^{t_i+\tau}$ is the average frequency over the interval $[t_i, t_i + \tau]$ and f_0 is the nominal or reference frequency of the synchronization signal. The fractional frequency deviation $\bar{y}_i$ represents the relative frequency error averaged over each interval τ , and it serves as the fundamental component in the Allan variance calculation.

The total number of these fractional frequency values, denoted as M , is determined by the total measurement duration and the length of the averaging interval τ . Specifically, M is the number of non-overlapping intervals of length τ that can be extracted from the measurement data. This relationship ensures that M directly depends on the chosen τ , as longer intervals τ will lead to fewer available intervals M .

There are several versions of the Allan variance, i.e., the square of the Allan deviation, that can distinguish between different types of phase noise and better describe the temporal stability according to the various noises present [194]. The most common among these is the original non-overlapping Allan variance, or two-sample variance, which is the standard measure of frequency stability in the time domain. It is defined as:

$$\sigma_y^2(\tau) = \frac{1}{2(M-1)} \sum_{i=1}^{M-1} (\bar{y}_{i+1} - \bar{y}_i)^2 \quad (\text{A.2})$$

where $\bar{y}_i$ is the i -th fractional frequency value, and M is the total number of such fractional frequency values that can be averaged over the interval τ .

The original Allan variance can be modified to make maximum use of a data set by forming all possible overlapping samples at each averaging time t and thus obtaining the so-called fully overlapped Allan variance. This index can be estimated from a set of M fractional frequency deviations for averaging time τ , by the expression:

$$\sigma_y^2(\tau) = \frac{1}{2m^2(M-2m+1)} \sum_{j=1}^{M-2m+1} \left\{ \sum_{i=j}^{j+m-1} (\bar{y}_{i+m} - \bar{y}_i) \right\}^2 \quad (\text{A.3})$$

where $\tau = m\tau_0$, m is the integer averaging factor and τ_0 is the basic measurement interval, i.e., the shortest duration over which individual measurements are made, and it is determined by the sampling rate of the measurement system. Therefore, the averaging intervals τ are constructed by aggregating multiple τ_0 -intervals. This aggregation allows for analyzing frequency stability over varying timescales and provides flexibility in adapting the analysis to different types of noise or application-specific requirements.

Compared with Eq. (A.2), this formula has a higher computational complexity due to the internal sum, and for this reason it is rarely used for large data sets, especially in real-time application. However, by taking advantage of all available data due to the overlapping intervals, a reduction in statistical uncertainty and a more accurate estimate of signal stability for each τ time interval can be achieved. This advantage is particularly glaring for noisy signals [194].

Other variations of the Allan deviation, such as modified Allan deviation,

time Allan deviation and the Hadamard deviation, can be found in the literature but will not be presented here because they were not used in the thesis. For further discussion, please refer to the literature cited in this appendix.

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List of Figures

1.1	Typical schematic of traditional centralized and modern decentralized electric power grids [14].	3
1.2	Diagram of a generic WAMPAC system, composed of the electric power grid (for instance IEEE 14-bus system [19]), instrument transformers, PMUs, PDCs, control center and actuator.	6
1.3	Typical connection diagram between transformers and PMU [24].	7
2.1	Example of the PMU Std for the amplitude measurement and the TVE of a 10 % amplitude step at $t = 0$	26
2.2	Typical block diagram of a standalone PMU.	30
2.3	Flow chart of the proposed TFM _{LR} algorithm.	38
2.4	Flow chart of the proposed TFM _{WRLR} algorithm.	42
2.5	Weight coefficients (components of $\mathbf{w}_{LR}$) parametrized by $-1 \leq \lambda \leq 0$	43
2.6	Magnitude response of the family of filters that enable estimating $\bar{X}_1$ parametrized by $-1 \leq \lambda \leq 0$	44
2.7	Phase modulation test: TVE as function of modulation frequency f_m as produced by TFM and TFM _{LR} in red and purple, respectively.	47
2.8	Maximum TVE for TFM and TFM _{WRLR} algorithms with different SNRs and sampling frequencies.	50
2.9	Amplitude step test in noiseless conditions. TVE as function of time as produced by TFM and TFM _{LR} in red and purple, respectively (a). Flag for left, right or full TFM expansion (b). .	52
2.10	Amplitude step test with SNR = 80 dB. TVE as function of time as produced by TFM and TFM _{LR} in red and purple, respectively (a). Flag for left, right or full TFM expansion (b).	52
2.11	Phase step test in noiseless conditions. TVE as function of time as produced by TFM and TFM _{LR} in red and purple, respectively (a). Flag for left, right or full TFM expansion (b).	53

2.12	Phase step test with $\text{SNR} = 80$ dB. TVE as function of time as produced by TFM and TFM_{LR} in red and purple, respectively (a). Flag for left, right or full TFM expansion (b).	54
2.13	Phase step test in noiseless conditions. Time evolution of TVE for TFM (solid red), TFM_{WRLR} (dash-dotted yellow), TFM_{R} (dash-dotted green) and TFM_{L} (dash-dotted blue) algorithms (a). Time evolution of λ (b).	55
2.14	Amplitude step test with 80 dB SNR. Time evolution of FE (semilog scale) for TFM (solid red) and TFM_{WRLR} (dash-dotted yellow) algorithms (a). Time evolution of λ (b).	57
2.15	Phase step test with 80 dB SNR. Time evolution of RFE (semilog scale) for TFM (solid red) and TFM_{WRLR} (dash-dotted yellow) algorithms (a). Time evolution of λ (b).	58
2.16	Amplitude step test with 72 dB SNR, $f_s = 50$ kHz. Time evolution of RFE (semilog scale) for TFM (solid red) and TFM_{WRLR} (dash-dotted yellow) algorithms (a). Time evolution of λ (b). . .	58
3.1	Exchange of synchronization messages between a master and a slave clock in IEEE 1588 [35].	70
3.2	Typical architectures based on local level synchronization with GNSS on the left and on a single wide-area synchronization source with WR technology on the right.	74
3.3	Schematics of the experimental test architectures implemented in the Cagliari laboratory (a) and in the METAS laboratories (b). .	77
3.4	Block diagram of the software implemented to synchronize NI cRIO devices with PPS.	79
3.5	Offset measurements as function of time in the considered cases.	80
3.6	Histogram of the offset measurements in the considered cases. . .	81
3.7	Quantile-quantile plot of the offset measurements in the four considered case.	82
3.8	Allan deviation of the offset measurements as function of integration time in the considered cases.	83
3.9	Boxplot of the phase-angle error due to synchronization as a function of the harmonic order for the WR Module case.	85
3.10	Block Diagram of a TW-based fault location system along a transmission line.	86
4.1	Comparison between architectures of traditional digital substation and future fully digital substation introduced by the standard IEC 61850.	90

4.2	Measurement chain for synchrophasor measurements based on SV inputs, as per IEC/IEEE 60255-118-1 [118].	92
4.3	Block diagram of proposed SV-PMU with time quality evaluation functionality.	95
4.4	Measurement setup for reproducing the synchronization loss at SAMU level.	98
4.5	Symmetric uncertainty ($k = 2$) affecting the SV-PMU publishing periods measurements as a function of nominal publishing periods t_{p0} ranging from 245 to 255 μ s.	100
4.6	Distribution of published (orange) and received (blue) SV, with a sinusoidal deviation w.r.t. nominal publishing period of 250 μ s realized with the following amplitude and frequency parameters: (a) $A = 10 \mu$ s and $f = 6$ Hz, (b) $A = 10 \mu$ s and $f = 4$ Hz, (c) $A = 20 \mu$ s and $f = 6$ Hz, (d) $A = 20 \mu$ s and $f = 4$ Hz.	101
4.7	Top plot: synchronization offset of the FPGA clock measured by the NI 9467 synchronization board (solid line) compared against t_p^{os} as measured by the SV-PMU (dots) vs time. Bottom plot: difference between measurements.	103
4.8	t_p^{os} as measured by the SV-PMU (dots) vs time and RT time quality evaluation algorithm values (solid line). Dashed line represents the time quality threshold of 10 μ s.	105
5.1	Diagram of the considered architecture of a generic WAMPAC system, composed of PMUs, WAMPAC Server and Control Actuator, with the associated latencies.	108
5.2	Schematic diagram of the laboratory characterization architecture.	112
5.3	Block diagram of the proposed WAMPAC Characterizer.	115
5.4	Block diagram of the WAMPAC Emulator.	119
5.5	Picture of the used hardware and connections for the laboratory characterization.	120
5.6	Schematic diagram of the experimental test architecture (a), integration of the network emulator (b), and of a commercial PMU (c).	122
5.7	Diagram of the test architecture for the Italian TSO's WAMPAC prototype validated with the 4-20 mA WAMPAC Characterizer version.	124
5.8	Histograms of the WAMPAC latency measured for 15 minutes, considering 2, 5 and 10 PMUs, RR at 100 fps and UDP small packet size.	127

5.9	WAMPAC latency measured by RT and GP hardware, considering 2 PMUs, RR at 100 fps and UDP small packet size.	128
5.10	From the top to the bottom: WAMPAC latency (red dots), control packet latency (brown squares), max PMU communication latency (green crosses), and WAMPAC Emulator processing latency (blue diamonds), considering 10 PMUs, 50 fps and UDP medium packet size.	129
5.11	WAMPAC latency measured for 2 hours, considering 2 PMUs, RR at 50 fps and UDP medium packet size.	130
5.12	WAMPAC latency measured for 2 hours through the integration of the network emulator without extra delay, considering 2 PMUs, RR at 50 fps, and UDP medium packet size.	131
5.13	WAMPAC latency measured for 2 hours through the connection to OpenPDC, considering 2 PMUs, RR at 50 fps and UDP medium packet size.	132
5.14	WAMPAC latency measured for 2 hours through the connection to OpenPDC and network emulator with an additional delay of 18 ms, considering 2 PMUs, RR at 50 fps, and TCP medium packet size.	133
5.15	WAMPAC latency of the Terna's WAMPAC prototype based on 4-20 mA actuator technology measured using the 4-20 mA WAMPAC Characterizer version for 2 hours, considering 2 PMUs, RR at 50 fps and TCP medium packet size.	134
5.16	WAMPAC latency of the Terna's WAMPAC prototype based on 4-20 mA actuator technology measured using the 4-20 mA WAMPAC Characterizer version for 10 hours with no time synchronization, considering 2 PMUs, RR at 50 fps and TCP medium packet size.	135

List of Tables

2.1	Response time requirements for amplitude and phase step tests of P- and M-class compliance	25
2.2	Accuracy requirements for steady-state and dynamic compliance tests of P- and M-class with $f_{RR} = 50$ Hz	27
2.3	Worst-Case Errors for P-Class Steady-State and Dynamic Compliance Tests (TFM and TFM _{LR} algorithms)	46
2.4	Worst-Case Errors for M-Class Steady-State and Dynamic Compliance Tests (TFM and TFM _{WRLR} algorithms)	49
2.5	Maximum and RMS Errors for Different SNRs in Nominal Conditions (TFM and TFM _{WRLR} algorithms)	50
2.6	Response Time of TVE, FE and RFE for TFM and TFM _{WRLR} in Linear Amplitude Transitions of Different Duration	59
2.7	Amplitude and Phase Step Response Times of the TVE for Different State-of-Art Algorithms	62
3.1	Time Synchronization Accuracy Classes for Power Systems Applications in IEC Standard 61850-5:2013	75
3.2	Statistical Features of the Offset	81
5.1	WAMPAC Latency Results for UDP Protocol and RT Hardware	126
5.2	WAMPAC Latency Results for TCP Protocol and RT Hardware	126
5.3	WAMPAC Latency Results for UDP Protocol and GP Hardware	128
5.4	Latency Components for 10 PMUs, 50 fps, Medium Packet Size, and UDP Protocol	129

Acronyms

A/D Analog-to-Digital

AC Alternating Current

ADC A/D Converter

AE Amplitude Error

AM Amplitude Modulation

ASDU Application Service Data Unit

BC Boundary Clock

BMC Best Master Clock

CERN European Organization for Nuclear Research

CESA Continental Europe Synchronous Area

CT Current Transformer

DC Direct Current

DER Distributed Energy Resource

DFT Discrete Fourier Transform

DGNSS Differential GNSS

EPRI Electric Power Research Institute

EV Electric Vehicle

FACTS Flexible Alternating Current Transmission System

FE Frequency Error

FIR	Finite Impulse Response
FPGA	Field Programmable Gate Array
FPS	Frames Per Second
GM	Grandmaster
GNSS	Global Navigation Satellite System
GOOSE	Generic Object Oriented Substation Event
GP	General Purpose
GPA	Grid Protection Alliance
GPS	Global Positioning System
HIL	Hardware-in-the-Loop
HT	Hilbert Transform
HVS	High Voltage System
IED	Intelligent Electronic Device
IpDFT	Interpolated DFT
IRIG	Inter-Range Instrumentation Group
IT	Instrument Transformer
LAN	Local Area Network
LFO	Low-Frequency Oscillation
LPIT	Low Power Instrument Transformer
LS	Least Squares
MCS	Master Control Station
MEO	Medium Earth Orbit
METAS	Swiss Federal Institute of Metrology
NCIT	Non-Conventional Instrument Transformer
NI	National Instruments

NTP	Network Time Protocol
OOB	Out-Of-Band
OS	Operating System
PAM	Phase-Angle Monitoring
PC	Personal Computer
PDC	Phasor Data Concentrator
PE	Phase Error
PLC	Power Line Communication
PLL	Phase Locked Loop
PM	Phase Modulation
PMU	Phasor Measurement Unit
PPS	Pulse Per Second
PQI	Power Quality Instrument
PTP	Precision Time Protocol
R-GOOSE	Rootable GOOSE
R-SV	Rootable Sampled Value
RES	Renewable Energy Source
RFE	ROCOF Error
RMS	Root Mean Square
ROCOF	Rate Of Change Of Frequency
RR	Reporting Rate
RT	Real-Time
RTDS	Real-Time Digital Simulator
RTT	Round Trip Time
S&H	Sample and Hold

SAMU	Stand Alone Merging Unit
SAR	Successive Approximation Register
SDG	Sustainable Development Goal
SFP	Small Form-factor Pluggable
SNR	Signal-to-Noise Ratio
STATCOM	Static Synchronous Compensator
SV	Sampled Value
SVC	Static Var Compensator
TCP	Transport Control Protocol
TF	Taylor-Fourier
TFF	TF filter
TFM	TF Multifrequency
THD	Total Harmonic Distortion
TIHD	Total Interharmonic Distortion
TQ	Time Quality
TSO	Transmission System Operator
TVE	Total Vector Error
TW	Traveling Wave
UDP	User Datagram Protocol
UTC	Coordinated Universal Time
VT	Voltage Transformer
WAMPAC	Wide Area Monitoring Protection and Control
WAMS	Wide Area Measurement System
WLS	Weighted LS
WR	White Rabbit
WS	WAMPAC Server

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"The future belongs to those who believe in the beauty of their dreams".