A Remote Carrier Synchronization Technique for Coherent Distributed Remote Sensing Systems

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Abstract—Phase, frequency, and time synchronization are crucial requirements for many applications, such as multi-static remote sensing and communication systems. Moreover, the synchronization solution becomes even more challenging when the nodes are orbiting or flying on airborne or spaceborne platforms. This paper compares the available technologies used for the synchronization and coordination of nodes in distributed remote sensing applications.

Additionally, this paper proposes a general system model and identifies preliminary guidelines and critical elements for implementing the synchronization mechanisms exploiting the inter-satellite communication link. The distributed phase synchronization loop introduced in this work deals with the self-interference in a full-duplex point to point scenario by transmitting two carriers at each node. All carriers appear with different frequency offsets around a central frequency, called the application central-frequency or the beamforming frequency. This work includes a detailed analysis of the proposed algorithm and the required simulations to verify its performance for different phase noise, AWGN, and Doppler shift scenarios.

Index Terms—Synchronization, multi-static remote sensing systems, distributed beamforming, Phase-Locked-Loops.

I. INTRODUCTION

Distributed payloads, decentralized systems, cooperative platforms, and collaborative beamforming are crucial elements enabling the next generation of multi-static remote sensing systems. Some examples of distributed remote sensing applications are the bistatic and multi-static SAR [1]. A recent example is the European Space Agency Harmony Mission (within the Earth Explorer 10 program), in which two identical receive-only spacecraft follow Sentinel-1D in a formation, and use it as a radar illuminator [2]. On the other hand, multistatic configurations are the only feasible alternatives to achieve the radar power budget in missions from MEO/GEO orbits [3]–[6].

Similarly, for microwave radiometry applications, the distributed and formation flying configurations would be a game-changing technology [7]. The spatial resolution of a single platform radiometer can be improved only by increasing its aperture size. Therefore, the use of formation flying configurations provides the potential to increase spatial resolution significantly [8]–[13]. This technique can also be applied in 3D synthetic aperture radiometers [14], [15], in contrast to 2D coplanar arrays, providing the system with more flexibility and giving the possibility to diminish the mutual coupling between the antennas. All these new configurations have stringent requirements in terms of absolute phase, frequency, and time synchronization. Since the signal generation is performed locally at each distributed node, achieving precise synchronization is a very challenging task. The synchronization becomes even more difficult when the geometric distance between the distributed nodes is considerable in terms of the signal wavelengths and, in particular, when this electrical distance is time-varying due to changes in the conditions of the transmission medium. This effect is observed for the case of nodes flying, hovering, or orbiting on aerial or space platforms.

The literature on distributed beamforming has been quickly populated during the last decade, and different theoretical and analytical models, algorithms, and techniques have emerged during the last years. Nevertheless, there are not many practical implementations, even at a research stage. The main limiting factor to make practical implementation feasible is the synchronization under realistic scenarios. In these cases, the requirements in terms of implementation effort, power consumption, and complexity needed to achieve the synchronization goals may surpass the ones required by the primary payload application itself.

Some practical solutions have been proposed in the mobile communications area, but the same technical challenges mentioned above were also an impediment for actual implementations. These impediments triggered the development of other Multiple-Input-Multiple-Output (MIMO) solutions with centralized synchronization relaying into reliability and trust (SnT), University of Luxembourg, 1855 Luxembourg City, Luxembourg (e-mail: juan.duncan@uni.lu; liz.martinez-marrero@uni.lu; jorge.querol@uni.lu; sumit.kumar@uni.lu; symeon.chatzinotas@uni.lu; bjorn.ottersten@uni.lu).

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backhauling-network links or GNSS signals instead of wireless synchronization techniques embedded in the communications standard. Besides, the synchronization backhauling via wires is not possible in all practical cases.

The field of Wireless Sensor Networks (WSN) has also attracted interest in distributed beamforming and remote clock synchronization due to rapid development in sensor technologies and embedded systems using low power equipment. Since WSN deals with weak power signals, the distributed and collaborative beamforming looks more appealing than in other types of wireless networks, in principle by the natural topology of the sensor networks, and also by its power constraints [16].

Several other application fields require synchronization of distributed radio systems, such as, for instance, very large baseline phase arrays. In the last years, several synchronization techniques for telecommunications using distributed radio systems have been proposed. In 1968, Thompson et al. [17] discussed, compared, and classified the available methods for reducing propagation-induced phase fluctuations in frequency distribution systems and defined the principle of round trip stabilization systems. These works were usually applied to very long baseline arrays for radio astronomy applications. This last case may also be seen as a distributed radio system, but with the nodes in a fixed static position. In these cases, the signal distribution and the remote synchronization is performed using an auxiliary wired or wireless transmission channel. The phase synchronization techniques described in [17] can be seen as a sort of distributed Phase Locked Loop (PLL). In those cases, a direct feedback action is applied over a two-way transmission media as part of a whole phase loop under the assumption of channel reciprocity. Even in this case, where the radio units have a fixed position and are interconnected with coaxial cables, the design is very challenging due to the persistent random variations of the electrical length of the transmission media used for the synchronization. These phase fluctuations come as a function of the temperature in the cables and other physical parameters that cannot be easily characterized. The synchronization becomes even more challenging when one or more nodes are constantly moving and a wireless media is used for the synchronization link. The main problems found in this case, on top of the mechanical movement of the nodes, are fading, multi-path and non-reciprocity of the channel, which makes a practical implementation of multi-static systems almost impossible in cluttered or indoor radio environments. Another challenge in distributed remote sensing systems is the trade-off between the radio resources, such as power and spectrum, used by the synchronization and coordination mechanism in comparison to the resources used by the sensing process itself.

Under these constraints, a set of scenarios for which distributed microwave remote sensing can flourish in future practical applications can be foreseen. The main factor that will benefit the implementation of distributed coherent sensing will be the availability of a cost-efficient inter-node communication channel suitable for synchronization and coordination. A good example of these scenarios is the one that comprises swarms or formation flying topologies implemented using spaceborne or airborne platforms.

In this paper, we study and compare the available technologies used for synchronization and coordination of nodes in distributed remote sensing applications. Additionally, we propose a dual-carrier remote phase synchronization system and identify preliminary guidelines and critical elements for the implementation of the synchronization mechanism.

This manuscript is organized as follows: Section II presents the system model to be used during the assessment. Section III contains a classification of the different synchronization systems. Section IV describes some of the examples of remote clock synchronization already present in the literature. Section V presents the drawbacks of the single-frequency in-band full-duplex synchronization loop. Section VI shows the analysis of the proposed dual carrier synchronization loop. Section VII shows the simulations of the dual carrier synchronization loop. Finally, Section VIII contains the conclusions of this manuscript.

II. SYSTEM MODEL

A typical remote sensing system consists of distributed wireless radio nodes. Nodes operate their remote sensing operations independently; however, they cooperate to perform beamforming operations. In general, the distributed remote sensing deployment can have different classifications according to the topology, nodes mobility, or how they perform the synchronization. The last one of them is the most interesting for this work. Specifically, for synchronization purposes, distributed remote sensing systems can be classified as follows:

- **Systems where the synchronization relies on the wired distribution of a common reference signal**, such as the Very Large Array (VLA) used in Radio astronomy.

- **Systems that depend on wireless communications to synchronize the distributed sensors**. For example, High Altitude Platform Systems (HAPS), stratospheric weather balloons, and satellite constellations. Our work is focused on this group.

The target applications for remote sensing, depending on the sensor type, can be either passive or active. Passive sensor nodes can perform only receive beamforming. However, with the active nodes, both transmit and receive beamforming can be performed. Nonetheless, both the active and passive nodes are equipped with transmit and receive modules in order to achieve synchronization, which is required to perform multi-beam beamforming.

In our system model, a distributed deployment of N radio nodes, i.e., no particular topology is considered. Each of them is associated with a state variable. Additionally, each node can communicate with other nodes in the array. All the variables are a function of time and frequency; hence, for the sake of simplicity, we do not use the time and frequency variables in the following representations. Let the channel matrix between the nodes denoted by \( H \), where \( h_{mn} \in H \) means the single tap channel for the waveform transmitted from the \( n^{th} \) radio towards the \( m^{th} \) radio. The channel response varies with time, altitude, and the radio frequency hardware. A typical schematic of the distributed sensing system is illustrated in Fig. 1.
A synchronization system for a distributed remote sensing application can be classified on the basis of radio resource allocation (frequency bands) for performing inter-node communication. There are three main classes, as described below and represented in Fig. 2:

- **Non-overlapping frequency bands**: In this type of system, various radio node pairs use orthogonal frequency bands for internode synchronization as well as remote sensing operation. Radio nodes operating on non-overlapping frequency bands can track the master reference signals. However, as the phase variations are not accurately tracked (due to significant gaps in the frequency bands), a divergence of the absolute phase is observed. Some notable examples of these systems can be found in [19]–[22].

- **Adjacent frequency bands**: In this type of system, the participating radio nodes use the same frequency band, but different central frequencies. In comparison to the non-overlapped frequency bands, the absolute phase variation at the frequencies of operation can be tracked accurately.

- **Overlapping frequency bands**: In this type of system, the radio nodes use the same frequency band and overlapping center frequencies. This type of deployment is capable of accurately tracking the phases at all the frequencies of operation. On the downside, in order to distinguish between the signals from multiple nodes, the signals need to be multiplexed in time or using other multi-user approaches. For example, in-band full-duplex or self-signal cancellation methods show satisfactory results under high signal to noise ratio (SNR) scenarios [23][24].

Fig. 1. Distributed set of radio nodes intended to perform a remote sensing system. The radio nodes are responsible for synchronization and beamforming. The elements $h_{11}$ to $h_{NN}$ are the self-interference elements that appear in an in-band full duplex channel interconnection. The self-interference signal reception can be used for self-calibration and self-tracking of parameters such as phase and timing offsets. $\mathbf{p}_n$ and $\mathbf{v}_n$ denote position and velocity of the nth node.

Since all the radio nodes are capable of autonomous operation, they generate their own initial time and phase reference signal. However, none of the references is equal to each other. The phases and drifts are also different to the extent that no coherent processing can be performed.

A synchronization among them is compulsory, the purpose of which is to establish a lock on initial time and phase reference signals with respect to a master reference. The master reference is selected to be the one with the best frequency stability and phase noise performance. To establish such a reference lock, every combination of the radio nodes $(n,m)$ transmit a waveform modulated with a known and controlled amplitude and phase [18].

Allowing every node to have a possible communication with other nodes will increase the SNR and, therefore, the quality of the synchronization of each node while avoiding the challenging requirements of big directive antennas and their accurate pointing. This mechanism will allow using all the elements in the array to disseminate the frequency reference given by the master node. These capabilities come at the cost of power and spectral resources. These resources can be split among the nodes in time, frequency, and code. For instance, if the DVB-S2X standard is used for this purpose, each node phase is only transmitted in a pilot field, and additionally, the waveforms from one node to the other nodes are multiplexed using different Walsh-Hadamard codes. Using this approach, an array of tens of nodes will require splitting the spectrum into tens of sub-carriers. Furtherly, for very dense arrays with hundreds or even more nodes, it will be reasonable that after a limited number of nodes (tens) operate in closed-loop synchronization, most of them synchronize passively to the array using the available signals.

### III. SYSTEM CLASSIFICATION

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### IV. LITERATURE EXAMPLES OF DISTRIBUTED REMOTE SENSING

The following list contains some relevant examples of distributed remote sensing systems and missions, not necessarily using the same synchronization approach presented in this work:

- **Tandem-X**: In the Tandem-X mission, which operates in X-band, two Synthetic Aperture Radar (SAR) satellites follow a close formation with the distance between them...
varying up to 100 meters [21][22]. Tandem-X is capable of performing its operation in both mono-static and bi-static configurations. In bi-static settings, where synchronization between the two satellites is required, one of the satellites operates as the transmitter while the other as the receiver. Further, a round trip synchronization procedure is performed to attain synchronization between the two satellites. A method for synchronization is detailed in [25].

- Low-frequency distributed radio telescope in space (OLFAR) [26][27]: An OLFAR system consists of a space-based low-frequency radio telescope which explores the “dark-ages” of the universe. The OLFAR system will consist of a minimum of fifty satellites and requires synchronization among them. The required synchronization is below one Degree; however, the frequency of target applications is centered at 30MHz.

- Laser Interferometer Space Antenna (LISA) [28]: In the LISA system, laser-based interferometry is used to detect the gravitational waves originating from galactic and extragalactic sources. By performing a coherent operation, a distance accuracy of approx. 20 ppm can be achieved.

- Gravity Recovery and Climate Experiment (GRACE) [29]: The objective of GRACE was to track changes in the Earth’s gravity field. This system consisted of two identical satellites in near-circular orbits separated by approximately 500 km while maintaining a synchronized inter-satellite link. The Gravity Recovery and Interior Laboratory (GRAIL) mission was an analogous mission, but in lunar orbit with two spacecrafts separated by 200 km [30]. In 2018, the GRACE Follow-On (GRACE-FO) mission was launched with very similar characteristics to its predecessor, but with a satellite separation of 220 km [31].

It is worth to mention that phase synchronization is also a harsh requirement for wired round-trip techniques used in radio astronomy applications (e.g. VLA [32]) and wireless techniques developed and applied in space Very Long Baseline Interferometry (VLBI) (e.g. VSOP [33] or Spektr-R [34]) where the coherent integration time may range from few minutes to hours.

V. ANALYSIS OF SINGLE FREQUENCY IN-BAND FULL-DUPEX SYNCHRONIZATION LOOP

The single frequency solution, which consist of two-transponders is represented in Fig. 3. The two satellites (nodes) work in a master and slave (follower) configuration, whereas the master has a high stability reference \( u_0(t) = e^{j\theta_0(t)} \), and the follower has a less accurate reference \( u_s(t) = e^{j\theta_s(t)} \).

To analyze the effect of non-ideal oscillators, we include the frequency and phase noise as the output of the oscillators \( u_m(t) = e^{j\theta_m(t)} \), and the reference \( u_s(t) \). The goal of the synchronization algorithm is to have the same phase at the beamforming clock reference in each node. This phase is not a static parameter, and it is affected by the communication channel too. For that reason, it is needed a distributed phase synchronization loop between both nodes to make the two phases \( \theta_{bf1}(t) \) and \( \theta_{bf2}(t) \) equal, despite the phase of the oscillator \( \theta_s(t) \) and the phase rotation introduced by the channel (The subscript of \( \theta_{bf1}(t) \) and \( \theta_{bf2}(t) \) is “bf” because these are the phases used for the beamforming operations). Notice that all the phases mentioned before are time-dependent. However, from now on, the variable \( t \) will be omitted of the equations for the sake of clearness.

The frequency response of the channel is represented in the scheme as its transfer function \( H(f) \) and its Laplace equivalent \( H(s) \). It is essential to note that, as a first approximation a single frequency and, hence, a symmetric channel is assumed.

![Fig. 3. Schematic diagram for the hypothetical single frequency in-band full-duplex synchronization loop.](image)

The output phase of the master node is determined by

\[
\theta_{bf1} = \theta_0 + G_m(s) \left( \theta_m - \theta_{bf2}H(s) - \theta_o - \theta_\theta \right),
\]

where \( \theta_{bf2} \) is the reference phase of the follower node, \( G_m(s) \) is the s-domain frequency response of the control loop in the master and \( \theta_o \) is the phase compensation required to make \( \theta_{bf1} = \theta_{bf2} \). As can be appreciated, the proposed algorithm is equivalent to a distributed PLL with acquisition and tracking stages determined by the loop equations.

Evaluating \( \theta_o = \frac{G_m(s)(2G_m(s)H(s) - \theta_\theta)}{2G_m(s)} \) and \( \theta_{bf2} = \theta_x(1 + G_x(s)) + \theta_{bf1}H(s)G_x(s) \) in (2):

\[
\theta_{bf1} = \theta_0F_{01}(s) + \theta_xF_{x1}(s) - \theta_mF_{m1}(s),
\]

where

\[
F_{01}(s) = \frac{G_m(s)}{G_m(s)(1 - 2H^2(s)G_x(s)) - 2}
\]

\[
F_{x1}(s) = \frac{2G_m(s)H(s)(1 + G_x(s))}{G_m(s)(1 - 2H^2(s)G_x(s)) - 2}
\]

\[
F_{m1}(s) = \frac{G_m(s)(2 + G_m(s))}{G_m(s)(1 - 2H^2(s)G_x(s)) - 2}
\]

Similarly:

\[
\theta_{bf2} = \theta_0F_{02}(s) + \theta_xF_{x2}(s) - \theta_mF_{m2}(s)
\]

With:
The distributed phase synchronization loop described by (32a) and (43a) allows \( \theta_{bf2} \) to track the changes in \( \theta_{bf1} \). However, there is a crucial limitation in this design. It is very challenging to implement the full-duplex link with the continuous transmission and reception of a single carrier frequency as is required for this solution. In an actual full-duplex implementation, there will be residual interference between the transmitted and received signal.

A solution to overcome these limitations is presented in the next Section.

VI. PROPOSED DUAL-CARRIER SYNCHRONIZATION LOOP

In the previous Section, the feasibility of a remote phase synchronization loop has been analyzed using the same carrier frequency. It is found that the main limitation for the implementation of the full-duplex loop is the ability to separate incoming and outgoing signals to detect the propagation phases and be able to compensate them.

Let us consider a hypothetical modified single-carrier scheme (per direction), as the one explained before, in which the forward and return waveforms use different central frequencies. In this case, the separation can be achieved by applying a frequency offset of \( f_{mo} \) to the return waveform. The follower node will generate this offset frequency using its own reference frequency. This frequency offset will inject to the loop a phase ramp of \( \theta_x (f_{mo} - f_m) / f_c \). On the other end, the master node will down-convert the received signal with a phase of \( \theta_0 (f_m - f_{mo}) / f_c \). Then, considering that the round trip steady-state phase gain is -1/2, the output phase of the complete loop will incur an unsolvable error of \(- (\theta_x - \theta_0)(f_{mo} - f_m) / 2f_c\). This phase offset, which varies over time, might be mitigated by reducing the difference \( f_{mo} - f_m \) up to the limit where the two carriers start to overlap. Additionally, the asymmetry in the round-trip loop (by using \( f_{mo} \) and \( f_m \)) will substantially augment the phase noise and phase drift injected by RF and microwave components in the loop and the phase offset of the transmission media. Therefore, symmetrical schemes would be preferable in order to address these challenges.

This Section describes a synchronization scheme between a master and a tracker (also known as follower or slave) using two frequency carriers per direction of propagation in a symmetrical fashion. In this scheme, the midpoint of the frequencies of the incoming signals is equal to the midpoint of the outgoing signals. Signal diversity by means of spread-spectrum orthogonal sequences could have been a solution. However, a prior temporal synchronization is necessary in order to cancel their mutual crosstalk, and this is considered out of the scope of this work. Fig. 5 shows a frequency plan of the

**DUAL CARRIER POINT-TO-POINT SYNC**

![Diagram of the proposed dual carrier synchronization loop](image)

Fig. 4. Schematic diagram for the proposed dual carrier synchronization loop. The dashed line represents the functions implemented digitally. Not shown in the scheme is the beamforming phase obtained as \( \theta_{bf} = \theta_{out} + \theta_x \). The purple blocks are the modulators. The demodulators are not explicitly drawn. The double compound lines represent complex variable (capable to express negative frequencies), and the single lines represent real variables.
proposed point-to-point synchronization scheme.

In the proposed dual-carrier scheme, the synchronization signals are placed at the edges of a central frequency $f_c$, which can be the central frequency used for the remote sensing operation. Fig. 5 shows a simplified block diagram of the proposed dual-carrier synchronization scheme. The scheme has the same objective of the single carrier synchronization loop explained in Section V, which is to transfer a very stable frequency reference from the master node toward a follower node.

The master node generates two reference frequency signals from an ultra-stable reference oscillator, which can be an atomic clock. One of these signals is the RF local oscillator used to upconvert the transmitted signals with the frequency $f_c$, and phase $\theta_o$ as shown in Fig. 4. The other signal is the frequency offset $f_m$ which is a complex sinusoidal generated in the digital domain. The master node transmits towards the follower node two modulated carriers, one with a frequency offset $-f_m$ with respect to the central frequency $f_c$, and another one with a frequency offset of $\alpha$ obtained from the whole compensation bidirectional loop.

The follower node receives these two signals after the channel propagation, with phases $\theta_{r1}$ and $\theta_{r2}$ in Fig. 4. The follower node uses an RF local oscillator with a frequency of $f_x$ trying to make it as close as possible to $f_c$ using a frequency source which can be much less stable than the master oscillator. In order to achieve this goal, the follower node demodulates the transmitted stream, and recover the impinging two phases, and then makes an average and tracks it with the transfer function $G_s$ to obtain the phase

$$\theta_{out} = \left(\frac{\theta_{r1} + \theta_{r2}}{2}\right) G_s$$

Fig. 4. Frequency plan for the dual-carrier point-to-point synchronization loop. The frequency $\pm f_m$ is the offset with respect to the main carrier $f_c$ of the synchronization signals sent by the master, similarly $\pm f_x$ is the offset of the return signals produced by the follower node.

This output is used to generate two modulated feedback waveforms which are transmitted back to the master node using the frequency offsets $-f_x$ and $f_x$. A similar process is performed in the master node to generate the loop compensation phase $\alpha$. The impinging modulated carriers, with phases $\theta_{r3}$ and $\theta_{r4}$ are demodulated, (this is not explicitly drawn in Fig. 4 for all the signals). The phases $\theta_{r3}$ and $\theta_{r4}$ of the two received modulated carriers are used to get

$$\alpha = \left(\theta_{r3} + \theta_{r4} - \frac{\theta_{offset}}{2}\right) G_m$$

Using this compensation phase the control loop cancels the propagation-induced phase fluctuation. As a consequence, the beamforming phase $\theta_{bf}$ at the follower node will tend to be equal to the phase of the master local oscillator $\theta_0$, with

$$\theta_{bf} = \theta_{out} + \theta_k \approx \theta_0$$

A. Transfer function analysis

A transfer function analysis is required to perform closed-loop design in the complete scheme since the output phase is controlled by the master phase and that the channel delays are compensated.

For the analysis, the channel is modeled by a delay $\tau$, idem for the four carriers and, a phase offset ($\theta_1, \theta_2, \theta_3, \theta_4$) which is the propagation phase wrapped in modulo $2\pi$.

Additionally, the received signal is assumed to be affected by an additive white Gaussian noise. These additive noises are assumed to be statistically independent since the carrier frequencies are different and non-overlapped. Some correlated effects may appear, for instance, due to a poor filtering of the LNA power supply. However, the consideration of these implementation aspects is out of the scope of the present work.

Fig. 6 depicts the channel model used for the phase compensation scheme.

The transfer function in the Laplace transform domain of the complete master node block is given by

$$G_s(s) = \frac{\alpha(s)}{\theta_{r3}(s)} = \frac{\alpha(s)}{\theta_{r4}(s)} = \frac{-0.5 G_m(s)}{1 - 0.5 G_m(s)}.$$  

This transfer function is used to obtain the phase transfer function from the master oscillator phase $\theta_0$ to the slave recovered phase reference $\theta_{out}$. This relation is found with the two signals transmitted to the follower node and retrieved in the phases $\theta_{r3}$ and $\theta_{r4}$. 

Fig. 5. Equivalent channel model for the dual-carrier phase synchronization scheme. The transfer function is modeled by a delay $e^{-\tau s}$ in the Laplace transform.

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![Diagram of channel model for dual-carrier phase synchronization](image-url)
\[
\frac{\theta_{\text{out}}(s)}{\theta_0(s)} \bigg|_{\theta_0=0} = \frac{H(s)G_c(s)}{1 - G_c(s)G_b(s)H^2(s)} \quad (9)
\]

On the other side, the transfer function from the phase of the follower node oscillator \(\theta_0\) to the \(\theta_{\text{out}}\) is

\[
\frac{\theta_{\text{out}}(s)}{\theta_0(s)} \bigg|_{\theta_0=0} = \frac{(H^2(s)G_c(s) - 1) G_b(s)}{1 - G_c(s)G_b(s)H^2(s)} \quad (10)
\]

Finally, the transfer function from the phases \(\theta_\alpha\) and \(\theta_0\) to the beamforming phase \(\theta_{bf}\) are found to be

\[
\frac{\theta_{bf}(s)}{\theta_0(s)} \bigg|_{\theta_0=0} = \frac{\theta_{\text{out}}(s)}{\theta_0(s)} \quad (11)
\]

and,

\[
\frac{\theta_{bf}(s)}{\theta_\alpha(s)} \bigg|_{\theta_0=0} = \frac{\theta_{\text{out}}(s)}{\theta_\alpha(s)} + 1 \quad (12)
\]

We propose to use second-order transfer functions for the PLLs of \(G_m(s)\) and \(G_c(s)\) because of known benefits, such as tracking capabilities and dynamic response. However, other kinds of loops may be further studied. The second-order transfer functions are described, as usual, by a natural oscillation frequency and a damping factor. Fig. 7-Fig. 9 show the Bode plots for the phase transfer functions in the complete system for the inputs \(\theta_0(s)\) and \(\theta_\alpha(s)\) for different natural frequencies and damping factors. For these plots \(\zeta_m\) is the damping factor of the loop \(G_m\) in the master and \(\zeta_s\) is the damping factor of the tracking loop \(G_c\) in the follower node. The channel delays are assumed to be negligible for the computation of the bode plots. However, these delays may be included in a stability analysis of the system.

**Fig. 6.** Frequency response for the synchronization loop for the \(\theta_0(s)\) and \(\theta_\alpha(s)\) inputs for different damping factors \(\zeta_m\) and \(\zeta_s\). Here the natural frequency in both the master and the follower is 200 Hz.

Fig. 7. Frequency response for the synchronization loop for the \(\theta_0(s)\) and \(\theta_\alpha(s)\) inputs for different damping factors. \(\zeta_m\) and \(\zeta_s\). Here the natural frequency in the master is \(\omega_m = 100\) Hz and the natural frequency in the follower is \(\omega_s = 400\) Hz.

**Fig. 8.** Frequency response for the synchronization loop for the \(\theta_0(s)\) and \(\theta_\alpha(s)\) inputs for different damping factors. \(\zeta_m\) and \(\zeta_s\). Here the natural frequency in the master is \(\omega_m = 400\) Hz and the natural frequency in the follower is \(\omega_s = 100\) Hz.

### B. Stability and Delay margin analysis

After finding the expression for the transfer function of the synchronization loop, the following step is to make a stability analysis to determine the scenarios and margins in which the system can operate. The system will be stable if all poles of the closed-loop transfer function remain in the left half of the s-plane. This stability is ensured if the magnitude of the open-loop transfer function \(G_c(s)G_b(s)H^2(s)\) fulfills the bode stability criterion [35].

The loop stability analysis does not have a gain margin since the open-loop gain is locked because all its factors have unitary magnitude. \(G_c(s)\) and \(G_b(s)\) are one-to-one tracking PLLs that are designed to be stable. The only external parameter that will have effects in the loop stability is the transport delay \(H(s)\). Fig. 10 shows a plot of the delay margin (the maximum acceptable delay before the system loss stability) as a function of the loop natural frequencies, \(\omega_m\) and \(\omega_s\) having that \(\omega_m = \omega_s\) for \(\zeta_m\) and \(\zeta_s\) equal to one. This plot shows that the delay margin decreases as the natural frequency increases. Here it can be seen that for the very atypical case of a natural frequency of 1 MHz, the maximum round-trip transport delay is \(\tau = 0.23\) μs, which corresponds to a distance of 34 meters in vacuum propagation.
Fig. 9. Delay margin for the synchronization loop as a function of the natural frequencies $\omega_m = \omega_s$. The plot is obtained for $\zeta_m$ and $\zeta_s$ equal to one.

Table I. Simulation parameters.

| Parameter             | Value                                      |
|-----------------------|--------------------------------------------|
| Standard              | DVB-S2X Super-Frame-Format 2               |
| Baudrate              | 8MHz                                       |
| Pilot duration (used) | 32 symbols (out of 36 in standard)         |
| Inter-Pilots period  | 956 symbols                                |
| Super-Frame duration  | 612540 symbols                             |
| Pilots modulation     | QPSK                                       |
| Pilot sequence type   | Walsh-Hadamard (on top of the superframe scrambler) |
| Master offset frequency $f_m$ | 50 MHz                                    |
| Follower offset frequency $f_s$ | 40 MHz                                    |
| Central carrier frequency $f_c$ | 2200 MHz                                  |

VII. SIMULATIONS OF DUAL-CARRIER SYNCHRONIZATION LOOP

We simulate the proposed dual-carrier synchronization loop using Matlab to validate it in a realistic scenario. The simulations implement the PLLs from the master and follower nodes and interconnect them using the scheme of Fig. 4.

Each PLL consists of a numerically-controlled oscillator, a phase discriminator, and a loop controller in the digital domain. The simulations use the numerology of a DVB-S2X [36] satellite communications using the super-frame format 2. The phase tracking loop in the master and in the follower is performed over the Pilot fields. As seen in Fig. 11, the Inter-Pilots period is 956 symbols.

Previous timing and coarse frequency acquisition in the receivers are assumed to be obtained from the frame structure. We made this selection due to the characteristic offered by the standard, such as the periodical pilot sequence used for synchronization. However, the proposed mechanism is general for any kind of coherent communication schemes. The objective of the simulations is to validate the model and verify that a very stable oscillator reference can be transferred to a remote node by canceling out the propagation-induced phase fluctuations. For this purpose, the simulation has the capability to include a phase noise mask for the oscillators in master and follower nodes.

Empirical models based on measurements suggest that the oscillator phase noise PSD can be described as a sum of power-law processes $h_\alpha f^\alpha$ with $\alpha \in \{-4, -3, -2, -1, 0\}$ [37] with an additional Gaussian segment near to the carrier [38]. However, for the sake of simplicity, the phase noise is modeled here by a two-state model proposed in [39], which includes a frequency walk, a phase-walk, and a white noise component. The accuracy of this approximation for low-cost oscillators was demonstrated in [40]. Fig. 12 shows a diagram of the model used to reproduce the phase noise behavior in the master and follower nodes.

Fig. 12. Estimation of phase noise spectrum in the RF oscillators at 2.2 GHz in the master node and the follower node. These responses are obtained from scaling-up two different qualities 10 MHz clock references. The spectrum is computed using the windowing described in Fig. 14.
The configuration parameters of the two-state oscillator phase noise model were selected according to the phase noise mask described in Table II.

Table II. Phase noise values for the master and follower reference oscillators at 10MHz.

| Phase Noise (dBc/Hz) | Master Node Oscillator | Follower Node Oscillator |
|----------------------|------------------------|--------------------------|
| At 10MHz             | 1 Hz                   | -85                      |
|                      | 10 Hz                  | -125                     |
|                      | 10 KHz                 | -160                     |

For the phase noise estimation, the samples are decimated by a factor of 956. A time series of around 501 seconds is used to obtain a total of $2^{22}$ decimated samples. The spectrum is obtained using the periodogram for 32 non-overlapped FFT blocks of $2^{17}$ samples. The master clock phase noise parameters are derived from the specifications of a commercially-available Low Noise Chip Scale Atomic Clock [41]. The follower node phase noise parameters assume any commercial-grade OCXO user in radio applications. The phase noise mask obtained in the RF oscillators shown in Fig. 13 is the reference oscillator multiplied by a 2200/10 scale.

The phase noise spectrum estimation is obtained by windowing the stream of samples by a Dolph-Chebyshev function shown in Fig. 14.

Fig. 13. Power response of the Dolph-Chebyshev window used for the phase noise visualization. The window has a duration of $2^{17}$ samples and it is designed to have a peak-to-sidelobe rejection of 300 dB.

The dynamic system described in Fig. 4 for the proposed dual-frequency synchronization system is simulated under realistic thermal noise conditions in the receivers. The signals in the receivers of the two nodes get to be $r_1' = r_1 + n_{r_1}$, $r_2' = r_2 + n_{r_2}$, $r_3' = r_3 + n_{r_3}$, and $r_4' = r_4 + n_{r_4}$, where $r_1, r_2, r_3$, and $r_4$ are the received signals, and $n_{r_1}, n_{r_2}, n_{r_3}$, and $n_{r_4}$, are assumed to be independent additive complex circularly symmetric white Gaussian noise random variables, with the same variance.

Under this simplification, it is assumed that the two nodes transmit the same power in the four carriers and that the four of them are affected by the same channel attenuation (this is a very realistic assumption since the channel is assumed to be reciprocal.). Therefore, we assume without any loss of generality that the SNR of the system is the SNR of the received signals in the master node and in the follower node.

The simulations are performed in the digital domain after the pulse compression decimation. Then, the decimated sampling frequency is $8\,\text{MHz}/956 = 8.36\,\text{kHz}$. The SNR of the received signal has a compression gain, from the 32 samples used in one Pilot field equivalent to 15 dB. The transfer functions in the decimated discrete domain are obtained from the digital implementation of the second-order PLLs, which consist of a loop controller with two integrators. The integrators are replaced by its discrete equivalent using the bilinear transformation $s = 0.5t_s(z + 1)/(z - 1)$. Where $t_s$ is the sampling frequency, and $z$ is the z-transform variable. The transfer function for the loop controller in the Laplace domain is

$$Y(s) = (2\zeta\omega + \omega^2/s)/s,$$

which is a function of generic damping factor $\zeta$ and natural frequency $\omega$. The phase discriminator is implemented with the “angle” function from Matlab and the Numerically Controlled Oscillator (NCO) with the exponential function.

Fig. 15 shows the phase noise for the $\theta_{bf}$ phase output in the follower node for different values of SNR and loop bandwidths $\omega_m$ equal to $\omega_s$. The damping factor for these plots is set to 1.
It can be seen how the phase noise of $\theta_{bf}$ tries to reach the phase of the master oscillator for low frequencies. It is observed that the contribution of the Additive White Gaussian Noise (AWGN) inside the noise bandwidth is equivalent to the SNR value, plus the 15 dB of compression gain, plus 6dB coming from the fact that the loop is dual-carrier.

The following figures assess the capabilities of the dual synchronization loop by looking into the difference $\theta_{bf} - \theta_0$ during a given time period, in this case, 120 seconds. Fig. 16 shows the values of this difference for a loop with 0dB SNR. Fig. 17 repeat the plots for a loop with 10dB SNR, and Fig. 18 does it for 20dB SNR.

Fig. 14. Estimated Phase noise spectrum for Beamforming phase $\theta_{bf}$, (which is the output of the dual carrier remote phase synchronization system) for different SNRs and loop bandwidth. Both damping factors $\zeta_m$ and $\zeta_s$ are set to one. The phase noise for $\theta_{bf}$ is estimated with the same procedure that for the one of the oscillators in the master and follower node, as is shown in Fig. 4. It is, 32 FFT blocks of $2^{17}$ samples using the window of Fig. 14.

Fig. 15. Time response of the phase difference ($\theta_{bf} - \theta_0$) for an SNR of 0dB and different loop bandwidths ($\omega_m = \omega_s$). The damping factors are fixed to one. The sampling rate is 8.36 kHz.
Fig. 16. Time response of the phase difference ($\theta_{bf} - \theta_0$) for an SNR of 10 dB and different loop bandwidths ($\omega_n = \omega_s$). The damping factors are fixed to one. The sampling rate is 8.36 kHz.

Fig. 17. Time response of the phase difference ($\theta_{bf} - \theta_0$) for an SNR of 20 dB and different loop bandwidths ($\omega_n = \omega_s$). The damping factors are fixed to one. The sampling rate is 8.36 kHz.

Fig. 18. Time response of the phase difference ($\theta_{bf} - \theta_0$) to an initial phase offset in the follower node oscillator of 180 Degrees. Two loop bandwidths are evaluated: 10 and 100 Hz ($\omega_n = \omega_s$). There is no AWGN in the simulations. The damping factor is 1.

Fig. 19. Time response of the phase difference ($\theta_{bf} - \theta_0$) to a frequency offset in the follower node oscillator of 50 Hz for two different loop bandwidths. There is no AWGN in the simulations. The damping factor is 1.

Fig. 20. Time response of the phase difference ($\theta_{bf} - \theta_0$) to a Doppler shift of 1 Hz due to a relative movement between the master and the follower node. Two plots are shown, one with an SNR of 10 dB and another one with infinite SNR. For both plots, the bandwidths are set to 100 Hz ($\omega_n = \omega_s$).

However, it can be observed that a severe Doppler can generate a constant phase offset. It can be roughly approximated to $\Delta \theta = 4\Delta f \cdot \xi_{m} \xi_{s} / \omega_m \omega_s$ radians, for a second-order tracking PLLs.

There is another effect produced by a relative movement between the two nodes, and it is a 90° phase ambiguity in the loop. Fig. 22 shows how the phase difference $\theta_{bf} - \theta_0$ jumps 90° when the phase of the frequency drift at 1 Hz reaches the 90° limit.
behaves as a function of the receiver noise, the variations in the phase noise of the follower oscillator, and to Doppler shifts. It is observed that for the Doppler shift or fast variations in the channel phase, the loop suffers a 90° ambiguity. These kinds of ambiguities are common in remote-phase synchronization loops, which usually contain frequency dividers. For this reason, future work should address the solution of the phase ambiguity in the received streams. It is worth to mention that the authors have an on-going hardware prototyping of the presented synchronization loop during the year 2020.

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