Vector Measurements for Wireless Network Devices

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Abstract

Wireless networks are an iconic technology of today’s modern era, they are present in our daily activities as can be exemplified by cellular communications, wi-fi, bluetooth, and others. Vector measurements play an important role in the design, simulation, and testing of wireless networks and are used to characterize key devices operating in the radio interface, such as amplifiers, filters, and mixers.

Accurate characterization is the key for improving the capacity and efficiency of wireless networks. As the demand for network capacity continuously increases, the accuracy of vector measurements must also improve. Further, it is anticipated that such trends will continue in the years to come. Consequently, the wireless industry needs to include nonlinear behavior in their characterization and analysis, to assess and guaranty the operation of the devices, and to comply to the specifications from governmental regulations. In contrast to linear behavior, nonlinear behavior presents an additional bandwidth requirement because the signal bandwidth grows when it passes through nonlinear devices. In this thesis, vector measurements for devices operating in wireless networks are studied, emphasizing a synthetic approach for the instrumentation. This approach enables the use of digital post-processing algorithms, which enhances the measurement accuracy and/or speed and can overcome hardware impairments. This thesis presents the design of a vectorial measurement system for wireless devices considering the aforementioned trends and requirements. It also explores the advantages of the proposed approach, describes its limitations, and discusses the digital signal processing algorithms used to reach its final functionality. Finally, measurement results of the proposed setup are presented, analyzed and compared to those of modern industrial instruments.
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To my family for giving love, support and encouraging me despite of the distance and the life waving. To my wife Pamela for all the happiness that she brought to my life.

Efrain Zenteno
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Part I

Introduction
Chapter 1

Introduction

Wireless networks are an iconic technology in today’s modern era. In recent years, wireless networks have experienced impressive growth, and this trend will likely continue in the years to come. Although the deployment of such networks has promoted the development of novel applications and scenarios; the hardware used to deploy these networks has become cheaper, and the capacity of the networks has increased.

Due to increasing demands from the users of wireless networks, operators struggle to offer greater capacity; thus, the importance of quick and accurate vector measurements is emphasized in modern industry. Lately, trends in “green technology” trend, which battle for the optimal use of energy, have enhanced the role of efficiency in wireless networks. The simultaneous demand for, capacity and efficiency, has shifted the characterization needs from linear towards nonlinear, and the present thesis discusses an approach for the required instrumentation.

1.1 Overview

To develop wireless networks, all of the different parts of the system must be understood, especially the hardware used in RF (radio frequencies), and the knowledge and understanding of various devices comes from measurements. The RF portion of the system is characterized using vector measurements. The term vector refers to the ability to simultaneously measure amplitude and phase of waves. In contrast, traditional microwave and radio measurements sometimes specify independent power or phase measurements.

Vector measurements provide engineers the knowledge required to design, deploy and maintain wireless networks. This measurements have a long and rich history that encompasses vector network analysis since the design of vector network analyzers (VNAs) to modern nonlinear network analyzers [1].


1.2 Contribution of the Thesis

The contributions of this thesis are presented in 4 papers based on synthetic instrumentation applied to vectorial radio measurements. These papers discuss the details of the setup design and post-processing algorithms required to achieve vector functionality using standard and generic hardware. The scenario posed in the setup includes the use of a single noncoherent receiver, which is widely available in test and workstations.

These papers exploit the flexibility gained by digitizing the baseband prior to up-conversion (as present in many modern instruments), which allows the use of digital signal processing (DSP) algorithms, from which the RF behavior can be extracted.

**Paper A:** Paper A presents the synthetic instrument used to perform vector measurements using a single receiver. This paper describes the procedure to maintain the phase reference required for simultaneous amplitude and phase measurement. This paper highlight the DSP algorithms to achieve the vector measurement and correction. Chapter 2 of the present thesis provides an overview of vector measurements for wireless network devices. Moreover, principles for synthetic instrumentation design are described, and their application in the design of the proposed synthetic vector network analyzer is briefly discussed in Chapter 3.

**Paper B:** Paper B proposes an extension of the synthetic instrument (SI) presented in paper A to perform nonlinear in-band measurements. this extension is possible considering that the measured bandwidth contains new frequency components in the measured band. Paper B addresses the calibration required to perform this type of measurements and introduces the power absolute calibration for the presented instrument, which the same excitation signal in the experiments is used as a time reference. Further discussion on this topic can be found in Chapter 3.

**Paper C:** Paper C summarizes paper A and paper B, and evaluates the speed, accuracy, and repeatability of the presented instrument (called SDM VNA). Further, paper C compares the SDM VNA with industrial modern vector network analyzers (VNAs) and describes its nonlinear abilities.

**Paper D:** Paper D describes a receiver bandwidth extension technique that allows measurement over larger bandwidths, which is required for nonlinear characterization. The proposed setup uses no additional hardware; facilitating the use of such schemes in operating tests and works stations. Chapter 2 of the present thesis describe the need for this type of measurement and explain the relationship between bandwidth, dynamic range and averaging. The suggested bandwidth extension suggested relies on the use of a known user-defined signal. Further, discussion on this technique can be found in Chapters 3 and 4.
1.2. CONTRIBUTION OF THE THESIS

Papers Included in the Thesis

The papers included in this thesis are listed below:


### 1.3 Acronyms

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Description</th>
</tr>
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<tbody>
<tr>
<td>AUT</td>
<td>Antenna Under Test</td>
</tr>
<tr>
<td>CW</td>
<td>Continuous Wave</td>
</tr>
<tr>
<td>dB</td>
<td>Decibels</td>
</tr>
<tr>
<td>DFT</td>
<td>Discrete Fourier Transform</td>
</tr>
<tr>
<td>DSP</td>
<td>Digital Signal Processing</td>
</tr>
<tr>
<td>DUT</td>
<td>Device Under Test</td>
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<tr>
<td>FPGA</td>
<td>Field Programmable Gate Array</td>
</tr>
<tr>
<td>LSE</td>
<td>Least Square Estimation</td>
</tr>
<tr>
<td>LSNA</td>
<td>Large Signal Network Analyzer</td>
</tr>
<tr>
<td>NVNA</td>
<td>Nonlinear Vector Network Analyzer</td>
</tr>
<tr>
<td>PAPR</td>
<td>Peak to Average Power Ratio</td>
</tr>
<tr>
<td>PC</td>
<td>Personal Computer</td>
</tr>
<tr>
<td>PSD</td>
<td>Power Spectral Density</td>
</tr>
<tr>
<td>RBW</td>
<td>Resolution Bandwidth Filter</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
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<tr>
<td>SDM</td>
<td>Software Defined Measurement</td>
</tr>
<tr>
<td>SI</td>
<td>Synthetic Instrument</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal to Noise Ratio</td>
</tr>
<tr>
<td>TOSM</td>
<td>Through-Open-Short-Match</td>
</tr>
<tr>
<td>USRP</td>
<td>Universal Software Radio Peripheral</td>
</tr>
<tr>
<td>VNA</td>
<td>Vector Network Analyzer</td>
</tr>
<tr>
<td>VI</td>
<td>Virtual Instrument</td>
</tr>
<tr>
<td>VSA</td>
<td>Vector Signal analyzer</td>
</tr>
<tr>
<td>VSG</td>
<td>Vector signal Generator</td>
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Chapter 2

Vector Measurements for Wireless Networks Devices

2.1 Overview

Digital components become cheaper, faster, and its flexibility has been enhanced a clear example of this trend is found in latest generations of programmed platforms as Field Programmable Field Arrays (FPGAs), which offer large computational power. The availability of such components is finding a place where only analog electronics were thought possible; this trend has enhanced the radio equipment gaining flexibility and allowing the use of generic hardware [2]. Hence, the abilities of the equipment are defined in the software; those called synthetic instruments [3, 4, 5, 6, 7] or virtual instruments [8], although the term synthetic seems to be gaining more followers.

Such combination is indeed the best of two worlds: generic hardware performing basic operations and digital capabilities allowing complex computations, corrections and different display formats. However, the implementation of the software in specialized platforms (as FPGAs) creates new challenges, since this devices are resource limited pushing for a branch of efficient algorithm implementation. This further trend creates what is defined as Software Defined Measurements (SDM) [9, 5].

2.2 Multisine Signals

Multisine signals became widely used in RF measurements; there are several reasons for its success:

- Multisines signals enable a simple mathematical framework where important figures of merit can be extracted both analytically and experimentally [10, 11, 12, 13, 14].
CHAPTER 2. VECTOR MEASUREMENTS FOR WIRELESS NETWORKS
DEVICES

• The digitization of baseband allows a simpler and quicker manipulation of these type of signals [15]. Further, multisine are periodic excitations hence its measurement can be performed avoiding leakage and other artifacts [16]; at the same time its PSD (power spectral density) is confined at discrete spots in the frequency plane enabling a high SNR (signal to noise ratio) in the measurement ensuring high measurement accuracy.

• Multisine signals can emulate a realistic excitations of the devices operating the wireless network [17]. They can mimic PSD [18] as well as higher order spectra [19, 20]. Hence, the nonlinear behavior obtained from such excitation is meaningful and may produce relevant figures of merit.

• Multisines are the choice for calibration in modern measurement systems as Large Signal Network Analyzers (LSNA) [1], Nonlinear Vector Network Analyzers (NVNA).

• Some effects in measurement system are well described in frequency, for instance, group and phase delay. Thus, the estimation of such parameters from multisines signals are robust since several frequency points are measured simultaneously.

2.3 Software Defined Measurements in Radio

Despite of the development of digital components, that follow approximately Moores’ law [21], the synthetic instrumentation has not followed in the same proportion. This effect is due to the slower development of components to serve as interface between RF and baseband [7], since they are key to enable SDM at RF.

Thus, The SDM path to RF can be tracked to the availability of cheap hardware components and integrated circuits which task is to transform the RF signals to low pass domain where digitization occur, and to the DSP algorithms developed to process, analyze, and display RF behavior. For instance, [4, 22, 23] distinguishes the physical layer as RF part, and leaves the baseband in digital stages which can be programmed. This trend can be evidenced in the so called Universal Software Radio Peripheral (USRP).

Another possibility is the direct use of DSO at the RF wave [24], this solution runs quickly into problems since Nyquist limit will require faster ADCs than today’s available in the market or which price become out of the budget for such instrument.

A great advantage of digital baseband is that digital processing algorithms can be applied to improve the system performance. Such improvements may tackle hardware impairments as IQ imbalance, transmitter nonlinearities, crosstalk, and others. A digital processing used with this purpose aligns with the so called “dirty radio” [25]. Many examples of this can be already found in the literature [26, 27]. The requirements for the design of instruments in the framework of SDM are analyzed and discussed [28]. Main points are the hardware generality and independence.
2.4. LINEAR AND NONLINEAR MEASUREMENTS

A synthetic instrument is depicted in Fig. 2.1. Three basic layers are distinguished: the physical layer, where the measurement occurs, the interface that controls the hardware, and the software layer that is basically a collection of DSP algorithms that are in charge of the processing, and presentation of the results.

![Figure 2.1: Layers of a synthetic instrument, hardware, interface and software.](image)

2.4 Linear and nonlinear measurements

Linear behavior is referred to the behavior of devices satisfying additivity, in other words, a device with transfer function $H \{ \cdot \}$ holding:

$$H \{ x + y \} = H \{ x \} + H \{ y \}. \quad (2.1)$$

Thus, nonlinear behavior is exhibit by devices violating eq. (2.1). This property allows to measure independently different spectral components of the linear response, as presented in Paper A and Paper C.

In contrast to linear operation, the nonlinear behavior is characterized by the creation of new spectral components. The class of nonlinear systems, studied in
this thesis, hold the property:

**Property:** A periodic excitation signal applied to a device exhibiting nonlinear behavior produces an output signal that has the same period as the excitation.

This property implies that the nonlinear behavior creates only new frequency components harmonically related, sub-harmonics can not be present in this type of systems. In principle, this property indicates that the time for the measurement of nonlinear behavior is not increased compared with linear measurement; since the period for the response remains the same either in linear or nonlinear behavior. However, the bandwidth becomes the limitation. **Paper D** contributes tackling the bandwidth problem, introducing the use of a known user-defined signal (called “pilot signal”). Such approach is similar to the phase calibration in NVNAs [1].

Today’s industrial nonlinear vector measurements use a reference signal for calibration [1] (called phase reference), which is calibrated by electro-optical system [29]. However, a nonlinear vector measurement is possible without the use of a phase calibrated reference, this has motivated some work [30]. But, this approach requires a calibrated (by nose to nose) digital sampling oscilloscope (DSO) together with coherent sources. Hence, the use of this setup becomes complex.

### 2.5 Continuous Wave and Modulation

The common excitation types to extract the response of a device are the Continuous wave (CW) and the modulated wave. In the case of linear systems, its intrinsic property (additivity) allows to use any of them without distinction. In contrast, finding the nonlinear response requires the use of a realistic signal, which is a modulated excitation.

However, the choice of excitation leads to a different measurement setup type, for instance, the receiver is narrowband for CW but not for modulated waveforms. This causes the dynamic range to be different.

The bandwidth of the receiver impacts the performance of the measurements. First, as the thermal noise (known also as Johnson-Nyquist [31, 32]) is directly proportional to the bandwidth of the measurement; thus, lower amount of noise is encountered in narrowband receivers. Secondly, as the Nyquist sampling condition is intrinsically related to the measurement bandwidth; thus, narrowband receivers require less sampling rate, which yields in general to better ADC performance. Hence, narrowband receivers offer the highest dynamic range. On the other hand, modulated measurements require a larger receiver bandwidth which reduces the dynamic range.

In contrast to CW excitation the modulated wave can excite multiple frequency lines simultaneously. Hence, the speed of the modulated measurement could be faster for dense frequency grids [33]. Further, the dynamic range in an instrument based on modulated signals may be enhanced by coherent averaging [34]. These
aspects are discussed in Paper C and depicted in Fig. 2.2 where three instruments are compared, the nonlinear vector network analyzer (NVNA) with a narrowband receiver architecture, the Digital Sampling Oscilloscope (DSO) with the largest receiver bandwidth and the SDM VNA (a synthetic instrument with an intermediate bandwidth).

![Figure 2.2: Comparison of three instruments for nonlinear systems with different architectures.](image-url)
Chapter 3

Synthetic Vector Network Analysis

This chapter presents a vector network analyzer implemented in the framework as a software defined measurements (SDM), called in the following SDM VNA, as described in papers A, B and C. The setup is formed by instruments already found in current test and workstations, where the ability to correctly represent vector quantities is performed in software in a PC.

We explore the hardware aspects of the design of such instrument, highlighting the compromises acquired for the use of generic or specific hardware. We describe the digital post-processing processes required to gain the vector measurement ability, relating them to the hardware design. Finally, we compare this synthetic instrument to a modern industrial VNA available in the market and describe some of this additional features gained by the synthetic design.

3.1 Setup

The proposed setup is formed by two RF instruments, currently found in test and workstations, a vector signal generator (VSG) and a vector signal analyzer (VSA), as depicted in Fig 3.1. A test set is built using directional couplers to separate the waves in different directions; a solid state RF switch multiplexes this four signals enabling a single receiver to perform the measurement.

Working principle

A VNA is capable of finding the complex ratio of two waves, such ratio includes amplitude and phase. The amplitude ratio indicates the attenuation / gain suffered by the waves, where the phase ratio indicates the time delay between them. A correct amplitude ratio may be found using any instrument with accurate power reading. However, the calculation of the phase ratio needs a simultaneous measurement of both waves, as done in current VNA architectures, a simultaneous measurement of different waves preserves what is referred as “phase reference”. A single receiver in
the SDM VNA architecture prevents for a simultaneous measurement of different waves. Thus, the solution, of the time multiplexing of the waves, does not keep the “phase reference”.

A key point to maintain the phase reference is that a single measurement of the time-multiplexed waves is long enough to record all waves multiplexed. Then the phase reference is kept in the VSA, despite of the measurement of different waves at different time instants. We later exploit the modulation capability of the VSG and VSA to recover the exact phase ratio, using the delay encountered in the reference wave.

3.2 Post Processing

The SDM approach includes the core of the measurement to the post-processing algorithms. This section describes these algorithms for achieving its functionality.

De-Multiplexing Signal

One task to achieve the VNA functionality is to separate (de-multiplex) all recorded waves from the recorded sequence. This procedure uses the power and the mean duration of the switching period as tools. The power of the recorded sequence is shown in Fig. 3.2, a slightly lower attenuation of the reference path compared to
3.2. POST PROCESSING

The rest of waves in the test set allows to use such levels to distinguish between different waves in the measured sequence.

The scheme for de-multiplexing enhances its robustness by using the duration of the switching period (denoted $\lambda$ in Fig. 3.2). Such addition helps to measure devices with both simultaneous high insertion loss and good return loss.

**Measurement system effects**

The excitation signal, is a multisine sequence described by:

$$x[n] = \sum_{k=0}^{T-1} a_k e^{j(\omega_k n + \varphi_k)}.$$  \hspace{1cm} (3.1)

This signal, composed by $T$ complex sines, is created in the baseband (low pass) form in a PC, uploaded to the VSG for up conversion and used to excite the DUT. A signal with flat amplitude $a_k = 1$ and phases $\varphi_k$ to minimize the peak to average power ratio (PAPR) is chosen [35].
CHAPTER 3. SYNTHETIC VECTOR NETWORK ANALYSIS

The output $z[n]$ of a linear system $H \{ \}$ excited by $x[n]$, can be written as:

$$z[n] = \sum_{k=0}^{T-1} H \left\{ a_k e^{j(\omega_k \Delta + \varphi_k + \Psi)} \right\},$$  \hspace{1cm} (3.2)

with a time delay $\Delta$, $\Psi$ as the phase delay introduced by the system. The magnitude response of the system $|H\{\omega_k\}|$ can be found by least square estimation (LSE) in:

$$z = A\theta + v,$$  \hspace{1cm} (3.3)

with $z$ as an $N$-sampled output vector on complex envelope form, $\theta$ as the vector containing the magnitudes and phases to be determined, and as $v$ is a vector representing imperfections from the measurement system approximated by a Gaussian process.

$$A = \begin{bmatrix}
1 & 1 & \ldots & 1 \\
e^{-j\omega_0} & e^{-j\omega_1} & \ldots & e^{-j\omega_{T-1}} \\
\vdots & \vdots & \ddots & \vdots \\
e^{-j\omega_0(N-1)} & e^{-j\omega_1(N-1)} & \ldots & e^{-j\omega_{T-1}(N-1)}
\end{bmatrix}.$$  \hspace{1cm} (3.4)

The solution to this equation is given by:

$$\hat{\theta} = (A^HA)^{-1}A^H z,$$  \hspace{1cm} (3.5)

with $^H$ as the hermitian operator.

The phase of this measurement $\angle \theta_k$ is:

$$\angle \theta_k = \omega_k \Delta + \varphi_k + \angle(H(\omega_k)) + \Psi.$$  \hspace{1cm} (3.6)

Under linear assumption and zero phase of the measurement system ($\angle \theta_k = 0$), the pseudo delay can be computed as:

$$\Delta = \frac{(\angle \theta_k - \angle \theta_{k+1}) - (\varphi_k - \varphi_{k+1})}{\omega_k - \omega_{k+1}}.$$  \hspace{1cm} (3.7)

Note that, the pseudo delay is the required delay to time align the multisine with its low pass excitation $x[n]$. Thus, $k$ in equation (3.7) can be chosen in the middle part of the band, where lower distortion from filters (generator or receiver) are expected.

Then, any phase in the measured signal can be compensated, time aligning it as:

$$\angle \theta_k^* = \angle \theta_k - \varphi_k - \omega_k \Delta.$$  \hspace{1cm} (3.8)

Both, the amplitude response from (3.5) and the pseudo delay from (3.7) are estimated using the measured wave corresponding to the excitation branch of the test set. However, both terms are used to compensate all the recorded waves.
3.3. RESULTS

Calibration

Calibration refers to the correction of the systematic deviations encountered in the measurement system. The calibration process uses the multisine excitation to take advantage and compute the frequency response measurement at all the excited tones, this is similar to [33, 36] with emphasis in time savings by this approach. Calibration is applied to the SDM VNA in order to obtain high accurate measurement, thus, the measurement procedure will indicate the DUT behavior at reference planes.

A 12 term error model have been used to perform the calibrations, such model include the following defects: imperfect source matching, lossy transmission and reflection tracking, limited isolation and directivity. Where the standards selected were: Through Open Short Match (TOSM). However any other type of calibration model and or standards is possible.

3.3 Results

The measured S parameters provided by the SDM VNA are presented below, the validation of the measurements is using a high-performance industrial VNA R&S ZVA-8.

![Figure 3.3](image)

Figure 3.3: Measured $S_{11}$ of an isolator. a) the magnitude of the measurement for the SDM VNA (solid) and the VNA R&S ZVA-8 (dashed). b) the phase difference for both measurement systems.
Fig. 3.3a) shows the measured magnitude $S_{11}$ for an isolator, in solid line is presented the VNA R&S ZVA-8, while in dashed lined the SDM VNA. Fig. 3.3b) plots the phase difference of these two instruments.

Fig. 3.4a) shows the measured magnitude $S_{21}$ for an 10 dB attenuator for both the VNA R&S ZVA-8 and the SDM VNA, in Fig. 3.4b) plots the phase difference of these two instruments.

As can be seen from Fig. 3.3 and Fig. 3.4 the measurements from these two instruments agree to large extend, in a later section an evaluation of the performance of this measurement is conducted.

The speed of the VNA is a feature of relevance, specially in test and work stations. In order to compare the speed of the SDM VNA with an industrial VNA, the settings of both instruments are chosen to have the same amount of noise when a through connection is measured. The setting that plays a mayor role in reducing the noise in the VNA is the IF bandwidth filter, while in the SDM VNA is the coherent averaging [37],[34] performed to the measured waves. In Fig. 3.5, the plotted pairs $(x, y)$ indicates the settings for both instruments to coincide in the level of measured noise.
3.3. RESULTS

The degree of agreement between the two measurement systems, namely VNA R&S ZVA-8 and the SDM VNA, is verified by means of vector difference. This vector difference is compared to the maximum bound by the noise in both instruments [38].

The vector difference $e$ is denoted as:

$$e = S^{VNA} - S^{SDM VNA},$$

with $S^A$ denoting a $S$ parameter measurement from the system $A$. The noise bound for the worst-case condition ($\xi$) is encountered when the noise contributions of both systems act constructively. As depicted in Fig. 3.6

$$\xi = |R_{VNA}| + |R_{SDMVNA}|. \quad (3.10)$$

$R_A$ corresponds to the worst-case deviation from the noise encountered in a measurement of the system $A$. Finally, the condition for a measurement agreement is that $e \leq \xi$.

Fig. 3.7 shows the maximum vector difference obtained from a set of measurements for both systems (called difference bound), and also the bound for the noise ($\xi$ also called sum of repeat boundaries).

Fig. 3.7 is an assessment of the quality of the measurements for the SDM VNA, as $e \leq \xi$ is satisfied. This analysis does not include other effects that can
Figure 3.6: Comparison of two vector measurements. a) Agreement $e \leq |R_A| + |R_B|$. b) Disagreement $e > |R_A| + |R_B|$

Figure 3.7: Verification of the proposed instrument. The worst-case noise condition (solid), the maximum difference found using a set of devices (dashed)

potentially affect the results, such as connector repeatability. If so, a statistical validation may be suitable; however, the test of repeatability boundary (Fig. 3.7) is stronger than any statistical test. Finally, we can conclude that the SDM VNA performs measurements of similar accuracy as the VNA R&S ZVA-8.
Nonlinear Abilities

A calibration modification is required to compensate for both amplitude and phase absolute calibration. The power absolute calibration is obtained by using a power sensor in the band of interest. The absolute phase calibration is not performed for this setup. However, for all in-band measured tones the phases preserve the correct relationships. The phases of the measured signal are compared to the user-defined phases of the excitation signal to estimate a delay (called for that reason pseudo delay). This is equivalent then to use the excitation signal as a reference for time aligning the measured waves.

It can be argued that an absolute phase calibration is not required as long as all the relative phases of all frequency components remain unchanged. An absolute calibration refers to the ability of indicating the phases at certain time instant. Usually, such instant is the zero crossing of the fundamental harmonic (zero phase of the fundamental). Despite of the use of this time instant is widely extended, there is no problem to use any other time as long as the phase relationships to all harmonics are preserved, some discussions on this topic is presented in Paper D.

This setup allows to measure calibrated waves at the ports of a device, for instance, the Inter-modulation Distortion (IMD) products at the output of an amplifier, as shown in Fig. 3.8 where a large change of the phase can be observed. These measurements are valuable to characterize this devices, to study memory effects and to provide data for modeling.

Figure 3.8: Phase of the upper third order intermodulation product in a two-tone test.
3.4 Bandwidth extension

Multisines presents advantages for the estimation of system effects as time delay and phase delay. Thus, the time delay estimation is made with reference to the “user defined” multisine [39]. Hence, multisines are used as reference signals in nonlinear instrument systems. Vector receivers as the SDM VNA, trade-off dynamic range for measurement bandwidth, as indicated in Chapter 2. Several efforts have been made to extend the bandwidth of a receiver that offer higher dynamic range, for instance, adding extra signals to determine the phases of the nonlinear terms [40, 41, 42]. However, such approaches where specifically tailored to two or three tone signals and have limited applicability. The idea presented in Paper D explore the bandwidth extension.

Approach

The basic idea to extend the bandwidth of a receiver is to concatenate adjacent spectral bands in a post processing technique as depicted in Fig. 3.9. However, to render the correct time domain signal, such concatenation requires to remove the measurement system effects as well as to keep the phase reference. We use a known signal called pilot to maintain the phase reference and to be able to remove the effects from the measurement system, specifically the time delay and phase delay. The addition of a signal in the RF path is a common practice that can be found largely in the field [40, 41, 42].

The model of a measurement $z[n]$ using a noncoherent receiver is denoted as:

$$Z[k] = X[k]e^{j(2\pi k/N\Delta + \Psi)},$$  \hspace{1cm} (3.11)

with $Z[k]$ as the DFT of the measured signal $z[n]$, $X[k]$ as the DFT of the measured signal $x[n]$, and $e^{j(2\pi k/N\Delta + \Psi)}$ as the distortion involved in the measurement. Such distortion is composed by:
3.4. BANDWIDTH EXTENSION

- Time delay: Introduced because the signal has been measured with a different reference point \((n = 0)\) with respect to the signal generation. This effect is described as the rotation of \(X[k]\) i.e. \(X[k]e^{j2\pi k/N\Delta}\). The quantity \(\Delta \in \mathbb{R}\) denotes the time delay, which is in general a non-integer.

- Phase delay: The carrier phase difference \(\Psi\) (between modulator and demodulator), affects the measured signal as \(X[k]e^{j\Psi}\).

An external generator with additional hardware can be used to insert this signal to the signal of interest [43, 44], but this will create distortion in the measurement. Thus, the addition of the pilot signal can be done in digital baseband as suggested in Fig. 3.10.

![Figure 3.10: Set up to add the pilot signal \(p[n]\) to the signal of interest \(x[n]\).](image)

Estimation of the measurement distortion

As mentioned earlier the estimation of both \(\Delta\) and \(\Psi\) is in the core of this technique. The estimation of \(\Delta\) is usually referred to as 'time delay estimation', and in RF two main approaches are distinguished: maximization of the cross-correlation function [45, 46, 47, 48], and a user-defined cost function [49, 39, 50]. However, these two methods are closely related, and in some cases, the estimates converge to the same value [51]. The estimation of \(\Psi\), called phase carrier recovery in telecommunications, can be performed using both hardware and software approaches [48]; an overview of these approaches can be found in [52].

The estimation of \(\Delta\) and \(\Psi\) is made using the separated pilot sequence \(\tilde{p}[n]\) (the measured \(z[n]\) corresponding to \(p[n]\)) and the known user-specified \(p[n]\). The phase
difference between the measured $\tilde{p}[n]$ and the known sequence can be expressed
using the DFT as:

$$G[k] = \angle \tilde{P}[k] - \angle P[k].$$  \hfill (3.12)

Note,

$$G[k] = 2\pi k \Delta + \Psi + \text{Im}(V[k]/Z[k])$$  \hfill (3.13)

with $V[k]$ the noise sequence, and Im represent the imaginary part of the complex argument. Such result is in agreement with the Tretter’s approximation [53].

With $E[k] = \text{Im}(V[k]/Z[k])$ as the noise contribution for the phase term. When $E[k]$ is normally distributed with zero mean $\sim \mathcal{N}(0, \sigma^2)$, (3.13) can be solved optimally by Linear Least Square Estimation (LSE) on:

$$\begin{bmatrix}
\omega_0 & 1 \\
\omega_1 & 1 \\
\vdots & \vdots \\
\omega_{T-1} & 1
\end{bmatrix}
\begin{bmatrix}
\Delta \\
\Psi
\end{bmatrix}
= 
\begin{bmatrix}
G[k_0] \\
G[k_1] \\
\vdots \\
G[k_{T-1}]
\end{bmatrix}$$  \hfill (3.14)

with $k_t$ as the frequency bin corresponding to $\omega_t$ the angular frequency of the $t$th tone. $T$ is the number of pilot tones around a center frequency. Thus, there are standard tools for this estimation that are robust and reliable. However, in practice, this estimator will lack robustness, since for large $\Delta$ the phase it is not linear but saw-tooth shaped due to phase wrapping. Hence, the estimation is enhanced by an initial sampled-based synchronization using cross correlation maximization [54, 45, 46, 55, 56].

**Verification**

The verification is made using a reference measurement with a bandwidth covering the whole signal of interest. First, a total band of 48 MHz will be measured by concatenating 6 adjacent spectral bands ($B^+ = 8$ MHz), such result is compared with a single measurement with the same receiver but with a bandwidth of 48 MHz (referred as reference measurement). Fig. 3.11 shows the power spectrum of the measured signals and the power spectrum of the difference. As can be seen, such differences are below 40 dB. Notice that the same level of differences can be encountered when the same signal is measured by two different instruments (two stand alone ADCs) as depicted for the measured phase in Fig. 3.12.
3.4. BANDWIDTH EXTENSION

Figure 3.11: The verification of the proposed bandwidth extension technique. The difference between the reconstructed signal and the direct measurement is below 40 dB.

Figure 3.12: Two measurements of the phases of the same multisine signal using two different stand alone ADCs.
Chapter 4

Discussions and Conclusions

4.1 Discussions

This thesis explores the design of a vector SI using the SDM (software defined measurement) paradigm. SDM can be viewed as the separation of the RF layer, which remains analog, and the baseband, which becomes digital. Thus, SDM benefits from the digital domain in the base band, enabling the enhancement of both hardware impairments and allowing the use of algorithms to overcome channel effects.

In the papers outlined in the present thesis, multisine signals are used as excitations to identify measurement system effects and enable suitable corrections. The presented setup is non-coherent introducing both group and phase delay to the measured wave (as described in Paper D). However, a phase delay does not alter linear ratios, such as S-parameters, which simplifies their calculation.

The SDM framework separates the analog RF from the digital baseband. The bandwidth available at the baseband (digital band) poses a limit for the bandwidth covered at the RF, which limits the overall system bandwidth. Fortunately, for linear systems, different frequencies are independent; thus, we can concatenate different bands to make a much larger measurement, which is the approach described in Paper A and Paper C. Alternatively, nonlinear systems do not retain the same property, and, different frequency components have an amplitude and phase relationship. Hence, any instrument designed using the SDM principle is limited to the receiver bandwidth, as described in Paper B. An attempt to overcome this problem is presented in Paper D. In the approach a known user-defined signal is used, which enables a bandwidth extension similar to the straightforward concatenation used for linear systems.

The proposed methodology presents the disadvantage of that the pilot signal must be known. A different approach may be to use compressive sensing algorithms [57]; however, this techniques remain under-develop for measurements in wireless networks.
4.2 Conclusions

This thesis presented the design, implementation and post processing algorithms required for a synthetic vector network instrument. The digital signal processing algorithms required to achieve a vector measurement functionality were emphasized.

Hardware limitations and the use of digital post processing to overcome the hardware impairments were discussed herein. Furthermore, the results were compared to those of modern industrial instruments, as the VNA R&S ZVA-8 showing good agreement with the proposed measurement system.

The proposed instrument, called SDM VNA, uses modulated signals to perform measurements; thus, the receiver bandwidth is larger compared to the CW receivers, which reduces the dynamic range further overcome by performing coherent averaging. A coherent averaging technique predicts an increasing gain for a larger number of averages; however, hardware imperfections limit this improvement.

The ability to select the pilot signal removes some of the limitations in current nonlinear phase calibration procedures, and favors measurements with narrow tone spacing (limited by the depth of the receiver memory), which can be needed it to study long term memory effects. In contrast, the use of this pilot signal could require a previous “characterization” measurement, such as performed for current phase references used in modern instrumentation, this is specially recommended for larger bandwidths where deviations from the low pass signal are expected.
Bibliography


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