A Synthetic Vector Network Analyzing Measurement System

Magnus Isaksson, Member, IEEE, and Efrain Zenteno, Member, IEEE

Abstract—In this paper, a synthetic vector network analyzing measurement system is presented. The system is based on a hardware setup, including a signal generator and a vector signal analyzer, with the vector network analyzing functionality implemented in the software. The measurements of the proposed system demonstrated comparable performance in terms of accuracy and speed compared with a modern traditional vector network analyzer, but it is more flexible due to its inherent software implementation. The proposed system’s ability to measure nonlinear phenomena is addressed and discussed, and some preliminary results are given.

Index Terms—Nonlinear, S-parameters, synthetic instrument (SI), vector network analyzer (VNA), virtual instrument (VI).

I. INTRODUCTION

Advances of software platforms that increase their capabilities and reduce cost have led instrument designers to develop new kinds of instruments, namely, the so-called virtual instruments or synthetic instruments (SIs), with applications in industrial and educational fields with promising results [1].

The reason behind the success of this type of instruments is that it presents good features regarding flexibility, modularity, hardware independence, generality, and cost [2].

Considering the requirements in many research laboratories and in industrial production test workstations, it has become necessary to create instruments with new capabilities that are able to change to the current test requirements [3]; that is what SIs can offer.

Any SI can be logically divided into layers, where tasks and functions can be defined at each layer, enhancing the modularity, generality, and flexibility and allowing interchangeability of hardware. The fact that SIs are basically software makes software-defined measurements (SDM) [4] an adequate terminology in this case.

For instruments in the radio frequency (RF) band, measuring the linear as well as nonlinear response, varying the excitation signal, and having flexibility and modularity in pre- and post-processing digital techniques in order to deliver measurements with accuracy and speed are some of the desirable features in the measurement system.

Fig. 1 shows the structure of typical vector network analyzer (VNA) architecture. A major disadvantage of this structure is the inability to measure excitation signals other than continuous wave (CW) signals due to a narrow band receiver, which is also depicted in Fig. 1. This prevents the use of other excitation signals with a higher bandwidth, which has become a necessity of modern device characterization and testing [5].

This paper presents an SI, with the ability to perform vector measurements, which will be addressed as SDM VNA for the rest of this paper. One of the differences of the proposed approach is the presence of only one receiver and the usage of modulated signals in the source; this will relax the requirements of synchronization and distortion in the receiver compared to any VNA architecture, as the one presented in Fig. 1. That is due to the fact that phase relationship is kept now in the low-pass equivalent signal, which is measured through a receiver with improvements on IQ impairments, and not in the RF carrier, where traditional VNA measures. The linear distortion introduced in the receiver will cancel off in ratio computation as in S-parameter measurement.
The hardware setup of the SDM VNA, presented in Fig. 2, is composed of a vector signal generator (VSG) as the source and a vector signal analyzer (VSA) as the receiver. The VSG can generate signals from single CW and multisine realistic and modern modulation schemes such as code division multiple access (CDMA), orthogonal frequency-division multiplexing, etc., providing a very good control of the excitation signal.

The VSA provides a higher bandwidth and an internal correction for the linear distortion appearing when a signal passes through its own resolution bandwidth filter; thus, the use of this receiver provides a sufficient environment to perform S-parameter measurements and presents good features to exploit and present nonlinear analysis; furthermore, this setup is formed by an equipment that is normally available in current workstations for production or test, showing a system with the same accuracy and similar speed as a traditional VNA but without the expense of such an instrument.

This paper describes the advantages of this proposed instrument, i.e., its design, analyzing issues as accuracy and repeatability, showing results of S-parameter measurements so as to compare them with modern industrial instruments, and discussing its nonlinear capabilities.

II. THEORY

The SDM VNA functionality was implemented in a software using a personal computer (PC). The S-parameter calculation is based on sampled data through a VSA as the hardware setup in Fig. 2 indicates. The recorded data are represented as the complex envelope of the RF signal sampled.

In general, a real RF signal with carrier frequency $f_c$

$$s(t) = \Re \{ [x(t) + jy(t)] e^{j2\pi f_c t} \}$$

(1)

can be represented with its quadrature phase components, namely, $x(t)$ and $y(t)$. The complex envelope is

$$s_i(t) = x(t) + jy(t)$$

(2)

and is, in general, complex valued. The recorded sampled version of the signal $s_i(nT)$ with the sampling frequency $f_s = (1/T)$ and the sampling period normalized $T = 1$ is referred to as $s(n)$ and is used throughout this paper.

The algorithm to calculate the final parameters includes the following six steps:
1) stimuli generation;
2) device under test (DUT) excitation;
3) response recording;
4) S-parameter calculation;
5) error correction.

These steps are described below.

A. Stimuli Generation

The stimuli signals used were multisine signals

$$u(n) = \sum_{k=0}^{M-1} a_k e^{j(\omega_k n + \varphi_k)}$$

(3)

where $a_k$ are the amplitudes and the angular frequencies $\{\omega_k\}$ are distinct; $-\pi \leq \omega_0 < \omega_1 < \cdots < \omega_{M-1} \leq \pi$, with $\omega_k = (2\pi f_k / f_s)$ where $f_k$ is the tone location in hertz relative to the carrier; and $f_s$ is the sampling rate in hertz. A signal with all the amplitudes $a_k = a$, all the tones equally spaced, and $M = 10$ were used in this paper; however, the number of tones is not limited by this number. The number of points of the signal created contains an exact number of periods, so that phase continuity holds after cyclical upconversion. The generality of the SDM VNA measurement system allows, however, the use of any stimuli signal. One difference from traditional VNAs using CWs is that the ratio of the peak to average power of a multisine signal is higher compared with that of a CW signal. The S-parameters are defined under linear conditions, which imply some practical constraints for the selection of the initial phases $\varphi_k$. The initial phases are chosen accordingly based from [6] providing lowest peak-to-average power ratio, which is the best choice for the phase selection.

Coverage of the VSA frequency band is achieved with the multisine signal; the complex-envelope signal constructed is fed to the signal generator via a local area network and upconverted to RF, as shown in Fig. 2.

The system bandwidth is limited by the signal analyzer bandwidth where 20 MHz was used, with internal instrument compensation for distortion; thus, the tone spacing $(\Delta f = f_{k+1} - f_k)$ of the stimuli signal created was calculated to cover the whole bandwidth, resulting in ten tones spaced by 2 MHz each; thus, in order to perform coherent averaging efficiently, the envelope period (coherent window) should be kept as low as possible to allow a good number of windows to be captured; the envelope period in samples of a multisine created with a sampling rate $f_s$ is defined by

$$E_p = \frac{f_s}{\gcd(|f_k|)}$$

(4)

where $\gcd(\cdot)$ represents the greatest common divisor operator; in order to minimize this equation, $\omega_1$ was chosen to be the integer multiples of each other; thus, the envelope period becomes

$$E_p = \frac{f_s}{\min(|f_k|)}$$

(5)

where $E_p$ is, indeed, the minimum since $\gcd(|f_k|) \leq \min(|f_k|)$. 
The dynamic range (DR) of the proposed SDM VNA is limited by the DR of the VSA, represented by the boundaries of distortion and noise floor; given that distortion will be avoided by properly selecting a reference level in this instrument, the noise floor will set the DR of the complete approach; to tackle this issue, coherent averaging is proposed in the time-domain sampled waveform; this will result in lower DRs than commercial VNAs.

The systems have to remain linear in order to perform accurate measurements; then, appropriate settings in both VSA and VSG are important to avoid distortion, i.e., reference level for the VSA and power level for the VSG; this is important for the measurement itself, because S-parameter measurement of active devices requires the right amount of power applied to the DUT to make sure that it is not under compression.

B. Digital Signal Separation

One of the key issues of the system is the digital separation algorithm. It refers to the ability to distinguish the different signal types (i.e., reference, reflected, and transmitted signals) separated by the use of the solid-state switch in the hardware setup. The switch was run totally independent of the other components in the setup to avoid any hardware dependence between the switch and the other units; as a drawback of this approach (nonhardware dependency), an extra processing stage (digital signal separation) needs to be included. However, this stage has proved to be very robust and simple in its implementation with computational low cost, leading to similar accuracy in commercial VNAs.

The proposed amplitude level (or the analogous power level) technique is based on the hardware construction in combination with signal processing. The hardware setup configuration ensures that the reference signal always has a higher power than the reflected and transmitted signals with some significant margin. The different signal types can be distinguished from the power level edges in combination with the period of the switching time extracted from the reference level signal type. In combination with the proper choice of coupling factors, an overall robust digital signal separation scheme is provided.

Fig. 3 shows an example of measured data. Each point in the figure represents the mean power of a window using coherent sampling over the ten-tone signal.

C. S-Parameter Calculation

The S-parameter calculation is divided into two steps: 1) relative phase and 2) power calculations. In 1), the system is compensated for the deviations in the phase from the expected linear behavior. The method was partly developed using ideas presented, e.g., in [7]–[9] and was built on the possibility of calculating the pseudodelay, τ_p, of the system. The method was also briefly described in [10]. The pseudodelay is calculated for the reference signal and is thereafter considered constant when applied to the other signals calculating the phase deviations. The assumption is valid as long as the interval is small enough to avoid dependence on clock or oscillator drift. The method was found to be accurate for the present paper. This assumption is good since a single measurement collects enough samples to let the switch pass through all its branches (reference, transmitted, and reflected), so the phase reference is kept by the sampler of the signal analyzer; in this context, the pseudodelay is equivalent to the time difference required to align the phases of the tones to its initial phases, as explained in [7], thus holding this value for all branches ensures the proper phase relation in any relative measurements.

Formulated on the complex envelope form, the system is excited with a multisine signal u(n), as shown in (3), with M tones with equal amplitudes, where the frequencies ω_k are relative to the RF carrier ω_c. The tones are equally spaced with tone spacing Δω; the amplitudes and phases of the different tones are found by solving

\[ y = H z + \xi \]  

where y is an N-sampled output vector in complex envelope form and z is an M-term vector holding the unknown amplitudes and phases. The term ξ models all the imperfections in the measurement system, which was considered a Gaussian distributed random variable. The matrix H has a Vandermonde structure

\[ H = \begin{bmatrix} 1 & e^{-j\omega_1} & \cdots & e^{-j\omega_{M-1}} \\ 1 & e^{-j\omega_1} & \cdots & e^{-j\omega_{M-1}} \\ \vdots & \vdots & \ddots & \vdots \\ e^{-j\omega_1(N-1)} & e^{-j\omega_1(N-1)} & \cdots & e^{-j\omega_{M-1}(N-1)} \end{bmatrix} \]  

and (6) is solved according to [11] as the minimum least square error

\[ \hat{\xi} = (H^H H)^{-1} H^H y \]  

where H^H denotes the complex conjugate transpose of H. The argument of z, i.e., arg[z], is dependent on several parameters, such as the measurement delay ϕ_d, the initial phase ϕ_k, arg[H(ω_k)], and the pseudodelay τ_p. For one of the tones k, a phase equation can be formulated as

\[ \arg[z_k] = \varphi_d + \varphi_k + \arg[H(\omega_k)] + \omega_k \tau_p. \]  

Assuming a linear behavior for arg[H(ω_k)] for the two tones in the frequency center, i.e., k = 5, 6, the pseudodelay τ_p can be
calculated according to
\[ \tau_p = \frac{\Delta \arg[z] - \Delta \varphi}{\Delta \omega} \] (10)
where \( \Delta \arg[z] \) and \( \Delta \varphi \) correspond to \( \arg[z_6] - \arg[z_5] \) and \( \varphi_6 - \varphi_5 \), respectively. The pseudodelay \( \tau_p \) is used to compensate the phase of the rest of the tones
\[ \theta_{k,\text{comp}} = \arg[z_k] - \varphi_k - \omega_k \tau_p \] (11)
with \( \theta_{k,\text{comp}} \) as the compensated phase. The phase calculations are first performed for the reference signal, and thereafter, the same pseudodelay is applied in the phase calculations of the reflected and transmitted signals to calculate \( S_{11} \) and \( S_{21} \).

The stimuli signal was constructed to be repetitive. The accuracy in the amplitude calculations was increased with the help of coherent averaging [12]. The technique is possible when high-precision time alignment for the different measured windows is achieved (see, for example, [13]). To further increase the accuracy of the measurements, the digital adjustment of the filter deviation for the in- and quadrature-phase components inside the VSA was used.

D. Errors and Calibration

The influence and treatment of random and systematic errors in the measurements are summarized as follows. Some of the possible random errors are the following: 1) thermal noise from the instruments and switch; 2) shot and flicker noise; 3) carrier frequency drift; 4) temperature drifts in the system; and 5) errors in instrument–computer communications. The random type errors are modeled as white Gaussian noise and minimized through the averaging procedure described earlier. The drift in 3) is very small, i.e., both of the instruments used provide very stable oscillators. The error in 3) was further minimized by using a relative frequency locking of the instruments. The error [4] was addressed by measuring at thermal equilibrium, i.e., the equipment was on for at least 30 min before the measurements were taken. The errors in 5) were not analyzed, but the communication through the local area network was documented to be of low-error nature and assumed to be negligible for the given application.

The systematic errors were addressed by the calibration scheme. The errors can be summarized by the mismatch \( (M_S, M_L) \), the tracking \( (T_R, T_T) \), the directivity, and isolation errors. The short–open–load–through (SOLT) 12-term model along with the one-path two-port calibrations was used in the experiments. The mismatch errors are due to the mismatches in the source and load; the tracking errors are due to variations when there are tracking signals in the reflected and forward paths; the directivity error arises mainly from imperfections of the couplers; and the isolation error represents the error caused by the signal leaking from one port to another without passing the DUT. The theory for the calibration technique is widely known, and a complete description of the problem with a detailed overview can be found, e.g., in [14], and a complete mathematical description can be found, e.g., in [15].

In the following, we present the results of the S-parameter measurements using the concept of an SDM VNA described earlier. The validation of the measurements was performed by comparison with high-performance industrial VNAs, i.e., HP8753D and R&S ZVA-8. Many measurements on both passive and active devices, along with repeatability, uncertainty, accuracy, and speed comparisons, were the foundation of the total property analyses of the proposed system. The calibration method was also a variable parameter in some of the analyses.

In order to achieve the results shown in this section, concatenated measurements were needed, so the total bandwidth 0.5 to 3 GHz was covered by the sweeping of 20-MHz band (around the carrier) of both the VSG and VSA. This whole band has a lower boundary limited to 0.5 GHz by the separation hardware used (poor response of the couplers for low frequencies) and to 3 GHz as the upper boundary being the top frequency in the VSA; however, the technique is not limited in any way to these frequencies.

As many measurement instruments, the SDM VNA shows a tradeoff between accuracy and speed; these features are related to the amount of samples recorded in each measurement band; this tradeoff is analyzed under the speed subsection.

A. Passive and Active Devices

In this section, some results based on the measurements on passive devices performed with the HP8753D are compared with measurements with the SDM VNA in the range from 0.5 to 3 GHz. The traditional VNA (HP-VNA) was a full two-port SOLT calibrated with an intermediate frequency (IF) bandwidth of 100 Hz, an IF averaging of 128, and 1601 points.

In Fig. 4, \( S_{21} \) measurements of a 10-dB attenuator are shown; Fig. 5(a) and Fig. 6(a) show the...
magnitude, and the phase is shown in Fig. 5(b) and Fig. 6(b); from both Figs. 5 and 6, a good agreement against the Hewlett-Packard (HP) VNA can be seen.

B. Repeatability Analysis

Here, the criterion given in [16] was used to perform the repeatability analysis. The method is based on the measurements of two consecutive calibrations under the same conditions. The changes in the systematic terms caused by the noise are used to estimate the deviation in any consecutive S-parameter measurement. The six systematic terms $\varepsilon_{\text{sys},i}$ considered are given earlier, and the maximum deviation is calculated as

$$D_R = \max \left\{ |\varepsilon_{\text{sys},i}^A - \varepsilon_{\text{sys},i}^B| \right\}$$

(12)

where $A$ and $B$ denote the two calibrations and $i = 1, 2, \ldots, 6$. The parameter $D_R$ defines the boundary for repeatability for the conventional and SDM analyzers. As an example, Fig. 7 depicts the analyzer from HP as the comparative analyzer. The figure shows $D_R$ on the $y$-axis. The comparison was performed for several averaging factors in the HP instrument, using 500 Hz of IF bandwidth; the narrower the IF bandwidth, the better the repeatability boundary (decreasing the slopes), as shown in Fig. 7.

The repeatability boundaries found using the HP and SDM analyzers are shown as functions of the frequency from 0.5 to 3 GHz. The slope of the repeatability bound for the SDM is dependent on the number of coherent averages performed to acquire measurements, as explained in the previous section in this setup.

Decreasing the number of windows in which coherent averaging is performed will lower the repeatability for the SDM VNA making this instrument comparable with lower averaging factors of the HP VNA; such change will lead to an increase in speed because less samples are to be captured; however, a good tradeoff is found for 30 windows as can be seen in Fig. 8, where it will indicate a comparable IF bandwidth of 100 Hz in a traditional VNA.

It can be readily found that, by using 128 as the averaging factor in the HP instrument, the SDM and the traditional analyzer are comparable.

In order to find the relation between the number of coherent averaging performed in SDM VNA with the VNA IF
Fig. 9. Comparison of repeatability between HP8753D and SDM VNAs.

bandwidth, the repeatability boundaries $D_{\text{r,SDM}}$ and $D_{\text{r,VNA}}$ are analyzed; in Fig. 8, the red squares show the VNA IF bandwidth compared to the number of coherent averaging windows required to achieve the same level of repeatability bound for both systems, expressed as $D_{\text{r,SDM}} = D_{\text{r,VNA}}$.

C. Accuracy Analysis

The accuracy assessment of the proposed instrument is based on the method presented in [16], and several devices were measured on both SDM VNA and traditional HP; the devices chosen were an isolator, an attenuator, a filter, and an RF power amplifier.

Using the results from the repeatability analysis, namely, $D_{\text{r,VNA}}$ and $D_{\text{r,SDM}}$, a maximum repeatability boundary (also called the sum of the repeatability boundary) can be expressed as

$$D_{\text{r}} = D_{\text{r,VNA}} + D_{\text{r,SDM}}. \quad (13)$$

Thus, $D_{\text{r}}$ represents the maximum allowed deviation in any measurement; from the set of devices, a maximum deviation in vector measurements can be calculated using

$$\delta_{ij} = \max \left( |S_{ij}^{\text{SDM}} - S_{ij}^{\text{VNA}}| \right) \quad (14)$$
$$\Delta D = \max (\delta_{ij}) \quad (15)$$

where $\Delta D$ represents the maximum deviation presented in the measurements of the chosen set of devices, also called the difference boundary.

Both instruments will perform equivalent measurements as long as the difference boundary $\Delta D$ is less than the sum of the repeatability boundary $D_{\text{r}}$, which can be expressed as

$$\Delta D < D_{\text{r}}. \quad (16)$$

Both limits are shown in Fig. 9 with respect to the frequency band used in the experiments. It should be clear that both instruments present a fairly good agreement for their respective measurements; however, because no connection repeatability is considered in this assessment method, great care in connector connections had to be taken to avoid larger variations, which can lead to poor results and even wrong conclusions.

D. Speed

In a typical VNA, the speed is proportional to the bandwidth of the filter used in the receiver, which is called the IF bandwidth, where, the larger the bandwidth, the faster the measurement. However, increasing this bandwidth introduces more noise, losing accuracy and reducing the DR.

In the SDM VNA, discussed in this paper, the speed is related to the processing time of the VSG and VSA in performing some commands, such as changing of carrier frequency and collecting a specific amount of samples in the receiver VSA. It was found, by experiments in this setup, that the “bottleneck” is caused by the delay in recording samples at the receiver.

Thus, SDM VNA holds the same relation as in VNA: “The slower the speed (more collected samples), the higher number of averaging is performed and so higher accuracy is achieved.”

To calculate the speed, ten consecutive measurements of the S parameters were taken from each instrument, and the time required to collect the data was measured, where the speed (in megahertz per second) is calculated as the ratio of measured bandwidth and the time required for that measurement to be completed. Every set of ten measurements presents a magnitude variance

$$\sigma = E \left\{ (|S_{ij}| - |\bar{S}_{ij}|)^2 \right\} \quad (17)$$

where $E$ denotes the expectation and $|\bar{S}_{ij}|$ denotes the average magnitude of the ten measurements; this variance is directly linked to the noise in the measurement itself.

In Fig. 10, the speed of both systems is presented against the variance found for each set of measurements; the VNA used in this experiment was R&S ZVA8 from Rohde & Schwarz.

From Fig. 10 and considering the slope of the two lines, the higher the speed, the higher variance the measurement will present; although the VNA increases the speed in higher rate than the SDM VNA, the latter presents higher speed for small variances, where accurate measurements are obtained; in other words, SDM VNA performs faster measurements than a VNA when highly accurate measurements are required.
Fig. 11. Error model, the reference, and measurement planes for the absolute calibration procedure; the (red) waveforms represent the measured quantities (subindex m) as well as (subindex d) the real quantities.

Fig. 12. Measured power of the upper third IM product as a function of tone spacing and source power for an RF power amplifier excited with a two-tone stimulus.

IV. DISCUSSION

Due to the modulation ability as well as the precise control and design of the excitation signal, accurate nonlinear measurements could be carried out, as proposed in [17], and therefore, the calibration procedure needed to be extended [18]–[20]. Compared to the standard VNA calibration, a new term was added, \( e_{01} \), as shown in Fig. 11; like all the error terms \( e_{ij} \), \( e_{01} \) is a complex number with magnitude and phase, and for this reason, the process to find is called absolute power and phase calibration. A detailed description of the process of absolute calibration for this setup can be found in [21].

The term “absolute” refers to the ability for deembedding waveforms at each reference port (plane) in the device under testing; all phases are related to the fundamental signal in a process called phase normalization. The waveform “a1d” is presented as zero phase and can be represented as a real number. In cases where the analysis involves intermodulation (IM) products, the phases of these spectral components are referred to the corresponding phases of the fundamental tones.

The facilities to perform frequency sweeps as well as the power sweeps provided by the VSG made it possible to characterize the power amplifier and to find operation points where the IM distortion value is minimum, as shown in Fig. 12, where the power of the upper third IM product is shown on the vertical axis; the average power supplied by the source and the tone spacing are shown horizontally.

The minimum value for distortion was found at 2 dBm of applied power with 650-kHz tone spacing; in Fig. 13, the phase, in degrees, of the same tone (three IM) is shown; it can be noticed from this figure that a change of 180° is produced along with this reduction in the power of the three IM.

A 62-tone signal with random amplitude and phase can be used as excitation, which resembles the characteristics of CDMA signal in the output spectrum and amplitude distribution, with a power amplifier as DUT; Fig. 14 shows the in-band amplitude and phase behavior of each tone, obtained by a different input power level (indicated in the right column); by this test, the in-band distortion of these signals can be identified.

A good feature of the proposed setup is that it reduces the impact of the effect of the phase variations for the use of a single receiver in the architecture; on the other hand, this architecture allows one to measure one wave at a time; thus, it can measure the responses of time-invariant devices, where harmonics are excluded due to the limited bandwidth of the VSA. Although this inability is a disadvantage in the architecture compared to other nonlinear instruments, such as large-signal network analyzer (LSNA), its use in power amplifier behavioral modeling compared to LSNA has been conducted, where results slightly favor this proposed system [22].
Hence, the setup presented shows good features for the characterization of linear and nonlinear devices, and it is a valuable tool for creating behavioral models as well as for studying memory effects.

V. CONCLUSION

A synthetic vector network analyzing measurement system has been presented. The system makes use of a novel software algorithm in combination with a specific hardware setup. Both the software and hardware parts have been explicitly described in this paper. The resulting system has the functionality of a VNA but is more flexible due to its inherent software implementation.

The proposed system has showed a comparable performance in terms of accuracy and speed with a modern traditional VNA. This paper has also included a repeatability analysis.

Finally, the ability of the system to measure nonlinear parameters has been discussed, and some measurements have been presented.

REFERENCES


Magnus Isaksson (S’98–M’07) received the M.Sc. degree in microwave engineering from the University of Gävle, Gävle, Sweden, in 2000, the Licentiate degree from Uppsala University, Uppsala, Sweden, in 2006, and the Ph.D. degree from the Royal Institute of Technology, Stockholm, Sweden, in 2007. He has worked on communication products with the Televerket, Sweden, during 1989–1999. He is a Teacher in signal processing for telecommunications and is currently the Head of the Department of Electronics, Mathematics, and Natural Sciences, University of Gävle. His main interests are in signal processing algorithms for radio frequency measurements and modeling of nonlinear microwave systems. He is the author or coauthor of many published peer-reviewed journal articles, books, and conference proceedings in the area and is currently the Head of research within the fields of electronics, mathematics, and natural sciences at the University of Gävle.

Efrain Zenteno (M’09) received the B.S. degree from San Agustin University, Arqueipa, Perú, in 2004 and the M.Sc. degree in electronics/telecommunications engineering from the University of Gávle, Gävle, Sweden, in 2008. He was a Teacher in the program of telecommunications engineering with the Universidad Católica San Pablo, Arqueipa. He is currently with the Center of Radio Frequency Measurement Technology, Department of Electronics, Mathematics, and Natural Sciences, University of Gävle, and with the Signal Processing Lab, Royal Institute of Technology, Stockholm, Sweden. His main interests are instrumentation, measurements, and signal processing algorithms.