Behavioral Model for Reducing the Complexity of Mixer Analysis and Design

A. Cidronali,[1] G. Loglio,[1] J. Jargon,[2] G. Manes[1]

[1] Department of Electronics and Telecommunications University of Florence, Florence 50139, Italy
[2] National Institute of Standards and Technology, Boulder, Colorado 80305, USA

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ABSTRACT: This article considers an approach for the behavioral modeling of the conversion mechanism in a nonlinear device suitable for the analysis of RF/microwave mixers. The core of the model consists of the conversion matrix of the nonlinear cell under investigation, which represents its linearization around the large-signal state. This approach allows for a straightforward implementation in Computer-Aided Design (CAD) using the conversion matrix constructed from either simulation or measurements, of which the latter is considered here. Model order is significantly reduced due to the absence of the local oscillator signal in the frequency analysis plan. The intermodulation products are calculated in relative amplitude and phase and allocated in the spectrum on the basis of the conversion matrix coefficients. We illustrate the technique by implementing this model in commercial CAD software, which allows an in-depth insight into the conversion mechanism and illustrates the application to the design of a subharmonic mixer.
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Keywords: behavioral modeling; conversion matrix; subharmonic mixers; nonlinear circuit design

I. INTRODUCTION

Modeling of microwave devices and circuits has seen a continuous effort directed at coping with the huge increase in complexity of circuits and systems. In the field of RF and microwave transceiver design, the task is becoming even more involved due to the complexity of the modern communication standards' physical layer, which allows several coding and digital modulation operative modes. This implies that analysis is required at different levels of the system hierarchy, mostly those operating with mixed analog-digital signals. Here, the effort is either to optimize each subsystem or to guess the best specification trade-off among the subsystems. With regards to the nonlinear circuit design, the modeling effort is aimed at concentrating the complexity of the device's nonlinear dynamics into a compact representation in which static nonlinearities and memory effects must be included. Usually the designer resorts to a behavioral approach when the device physics and operating mechanisms are not well understood, or when the device is so complex that a circuit description would be impractical.

This opens the door to behavioral modeling, which allows a compact and potentially accurate representation of either subcircuits or subsystems, which can provide an improved numerical efficiency and, in turn, a reduction in the complexity of CAD simulations. In co-simulation of base-band circuit at high frequencies, the designer does has not have much flexibility due to the different time scales involved in the analysis. Although the envelope method can mitigate the issue of the transceiver complexity or even a complete mixer, amplifier, or oscillator, it prevents the use of a circuit-level description of each single building block. In a behavioral model, the description

Correspondence to Alessandro Cidronali; e-mail: alessandro.cidronali@unifi.it.
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is generally provided in terms of state functions, which are vectorial multiple-input equations of the so-called state variables. They are conventionally determined by processing small-signal and DC measurements. Recently, new techniques based on Vectorial Large Signal (VLS) measurements have been proposed to identify and validate such a class of models in both the time and frequency domain [1, 2].

The model ideally possesses the following features [3]. The first consists of the simulation capability; that is, given a finite set of previously measured excitations and responses, the corresponding results, obtained by simulations, should represent a good approximation of such data. Moreover, the model should display the same qualitative behavior that the device is known to exhibit when connected to an appropriate excitation network. For example, if the physical device is known to give rise to intermodulation products under certain excitations, then the model should exhibit the same behavior. Finally the model should be capable of predicting previously unknown operating modes. Such features are of great relevance in the exploration of new experiments and applications. These attributes are featured in the following approach.

This article is aimed at introducing a comprehensive frequency domain model, its CAD implementation, and its use for reducing the complexity in mixer analysis and design. The model is based on a recently introduced technique that allows us to extract, directly from VLS measurements, the conversion matrix (CM) of a nonlinear device. The CM is then adopted as a behavioral representation of the linearized nonlinear device under the action of a local oscillator (LO). A Schottky diode-pair used for subharmonic mixing is selected as a case study for the discussion of this approach.

The article is organized as follows. Section II examines the admittance CM as a behavioral model of a generic nonlinear device driven in a large-signal regime, while section III introduces the experimental behavior of the nonlinear cell in the CAD environment. Finally, section IV reports the analysis data related the design of a subharmonic mixer based on the behavioral model.

II. ADMITTANCE CONVERSION MATRIX AS BEHAVIORAL MODEL OF A NONLINEAR DEVICE

Let us consider a nonlinear device (ND), driven by a large signal at frequency F_{LO}. The linearization of the ND considers a frequency difference between the large and the small signal, F_O, hence the resulting intermodulation products (IMPs) contain the frequencies $nF_{LO} \pm F_0$, with $n = 0, 1 \ldots N$, where N is the highest considered local oscillator harmonic. The equations relating the current and voltage phasors at IMPs are linear and they are ruled by: $\mathbf{I} = [\mathbf{Y}] \times \mathbf{V}$. The CM's coefficients, $y_{m,n}$, relate the current i_m at frequency $mF_{LO} + F_0$, to the corresponding voltage v_n at frequency $nF_{LO} + F_0$, with all the voltage components differing from v_n equal to zero.[1]

The symbol $[\mathbf{Y}]$ represents the CM, which is, in fact, a linearization of the ND, under the large-signal state, with respect to the small signals, while the $\mathbf{I}$ and $\mathbf{V}$ vectors contain the current and voltage phasors at all the IMPs. The CM is adopted to furnish the ND's conversion properties, their dependence on the intermediate frequency, and the optimum impedances, not only in the mixer analysis but also in the noise figure minimization [4], which can be effectively involved in the analysis of the phase noise of oscillators [5].

The extraction of $[\mathbf{Y}]$ can be obtained either from circuit simulations, once the equivalent circuit model for the ND is available [6], or, as in this case, from VLS measurements methods as discussed in detail in [7, 8]. The latter returns a complete and accurate representation of the behavior, in particular concerning the nonlinear memory effects that give rise to nonsymmetrical output spectrum in transmitters. It is widely accepted that this latter feature is one of the critical issues to be managed in any up-converter design.

The objective of the study deals with the implementation of $[\mathbf{Y}]$ in a commercial CAD package and its use in the analysis and design of a subharmonic mixer.

Figure 1 shows the simplified schematic representation of the proposed model implemented in a CAD environment, including the required components, which consist of a frequency-defined device (FDD), an ideal two-port equation defined linear device (S2P_Eqn), and a linear S-parameter multiport (SNP) device. The number of ports required for the latter depends on the highest LO harmonic considered. The model is based on the conceptual link between the IMP's frequency and the virtual ports of the ND. The coefficients of the CM are stored in the SNP regardless of their spectrum allocation, which means that the SNP contains only the raw data referenced to an arbitrary frequency; e.g., the intermediate frequency, F_{IF}. Hence, the SNP converts the behavior associated with a specific CM coefficient, as the admittance seen

[1] For the lower side intermodulation products, the coefficients $y_{-m,-n}$, relate the current at frequency $mF_{LO} - F_0$, and the corresponding voltage at frequency $nF_{LO} - F_0$.

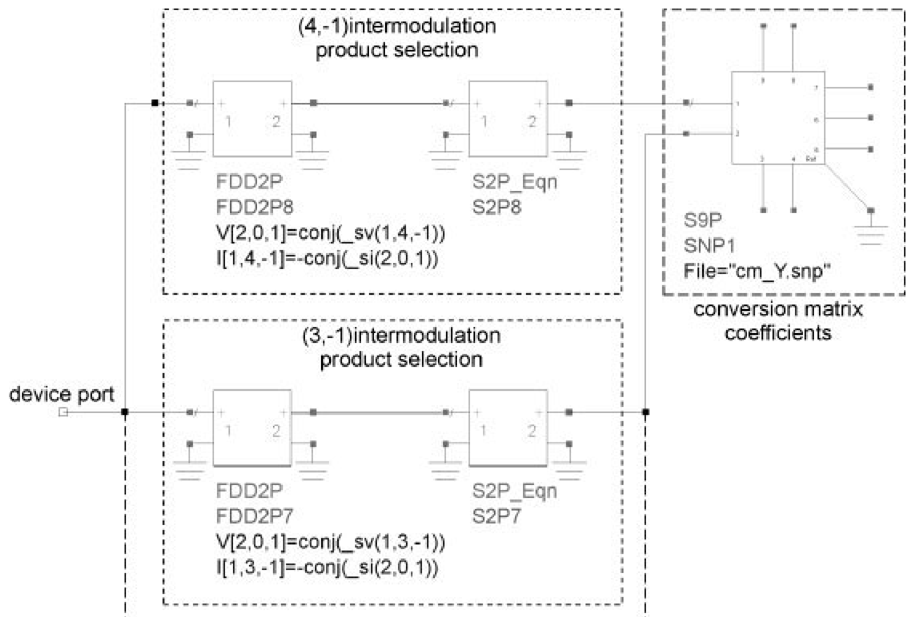


Figure 1. Schematic representation of the implementation in a commercial CAD of the frequency-domain behavioral model based on the CM.

at the relative port when all the other ports are shorted. This linear block is interfaced with the remaining circuit by an ideal frequency converter, which acts, for any port, on the appropriate phasor. The operation is schematized as follows:

- the voltage phasor is sensed at the FDD port 1 and converted in a voltage phasor at the fixed frequency at port 2, in the case under consideration F_{IF};
- the voltage phasor is applied to an ideal network, the S2P_Eqn, that has the following definition:
  – S12 = S21 = 1 if freq $=F_{IF}$, 0 elsewhere;
  – S11 = S22 = 0 if freq = F_{IF}, 1 elsewhere;
- the current phasor is originated by the applied voltage at the specific port of the SNP and by all the other port voltages present, at the same frequency;
- the above current phasor is then converted back by the FDD to its port 1 at the appropriate frequency.

The flow described above permits us to analyze a network operating in a multifrequency mode, like the case of an HB simulation, by concentrating all the information about the intermodulation generation in the linear analysis of a multiport device combined with ideal frequency converter blocks. Note that neither the FDD nor the S2P_Eqn introduce any weighting of the data; their purpose consists rather of the conversion from the nonlinear analysis to the linear one and vice versa. Moreover, the HB analysis is now retained only for its compatibility with the FDD environment. The presence of the LO signal is no longer required, having been already considered implicitly in the CM definition. The intermodulation order is defined by the FDD specification, thus reducing the complexity of the analysis.

III. MEASURED ADMITTANCE CONVERSION MATRIX OF A SCHOTTKY DIODE-PAIR

The method explained in section II is applied to implement the CM of two commercially available, general-purpose Schottky diodes, packaged in the same die, and connected in series with access to the central knot. The experimental extraction set up for the CM [7] considers the frequency plan: $f_{LO} = 2.6$ GHz, power ranging from –5 to +10 dBm, and a small tone $f_n = nf_{LO} \pm f_{IF}$, with $f_{IF} = 650$ MHz, –40 dBm power. The CM is then calculated by using the method in [8]. Henceforth, the experiments are relative to the diode-pair in the shunt configuration. Figure 2 reports the results of the CAD analysis related to the voltages and currents at the diode-pair termination as they appear in a CAD data display, for an 8 dBm LO level and a 0 dBm test signal injected at 650 MHz. It is easy to see the absence of the LO signal and its harmonics for the above discussion, while it is also noticeable the quantitative response of the device to

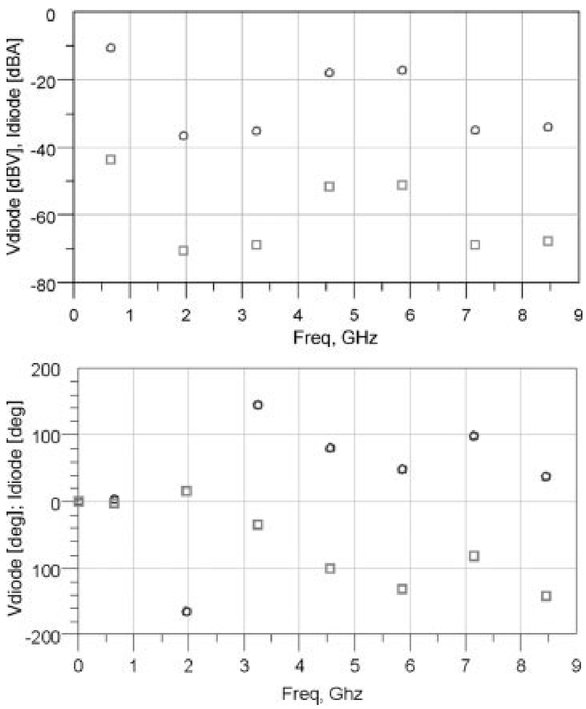


Figure 2. Linearized response of the device to a 0 dBm signal injected at 650 MHz, for 8 dBm LO level. Voltages (circles), currents (squares).

the 0 dBm test signal injected at 650 MHz. All the expected IMPs are present, particularly the third-order ones, namely, 2LO $\pm$ IF, and the sensibly lower second-order, namely, LO $\pm$ IF, confirming the quality of the device as a nonlinear element to implement a subharmonic mixer. Analogous results are also observed in the down-conversion regime.

Once the data in the format represented by Figure 2 have been collected, the CM is directly calculated and the meaningful information at the device level is available through a simple single tone HB simulation. These data are parameterized by the LO power level, which defines a CM for each LO level. This is the only way the LO enters into the analysis. As an example, Figure 3 reports in the polar plane on the measured conversion coefficients:

$$\Gamma_{5,1} = b[5]/a[1];\ \Gamma_{1,5} = b[1]/a[5], \qquad (1)$$

as a function of the LO power, associated with the up- and down-conversion, respectively, due to the second LO harmonic and in the upper side band case. Here, *a* and *b[j]* represent the incident and the scattered waves, respectively, at the i-th and j-th IMP frequencies (see Fig. 2). From Figure 3, we see that the amplitudes of the converted signals are almost the same in the up- and down-conversion modes, which

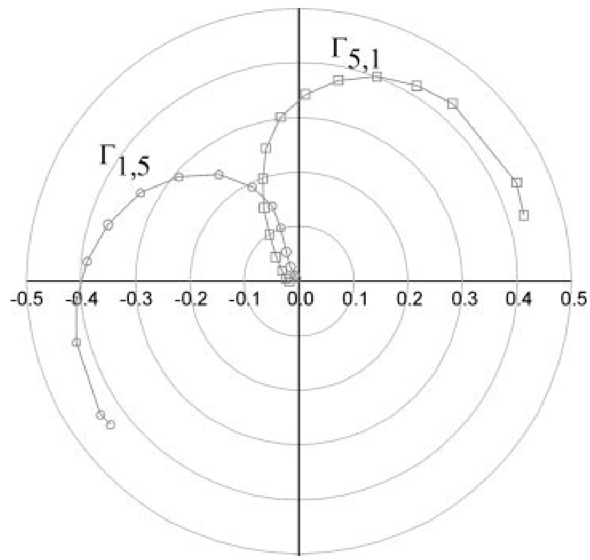


Figure 3. Measured Γ_{51} and Γ_{15} reflection coefficients, as functions of the LO power.

might be expected assuming that a subharmonic mixer, based on such a device, could be considered reciprocal. A closer inspection reveals that only at an LO power value, around +1 dBm, the conversion coefficients overlap, while at lower and higher values the reflection coefficients are slightly different in amplitude but overall exhibit a completely different phase behavior. The difference between the $\Gamma_{5,1}$ and $\Gamma_{1,5}$ conversion coefficients is due to the frequency of the injected voltage that is at IF and RF frequencies in the two cases, respectively. This, in turn, produces a

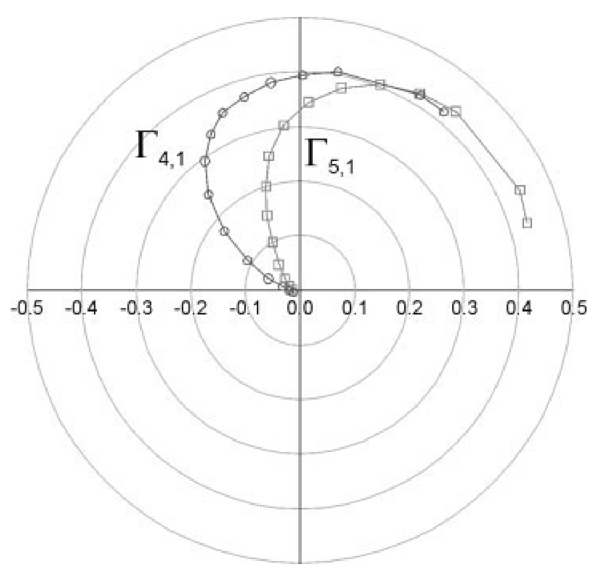


Figure 4. Measured conversion coefficients related to the subharmonic up conversion for the upper, $\Gamma_{5,1}$, and lower, $\Gamma_{4,1}$, side bands.

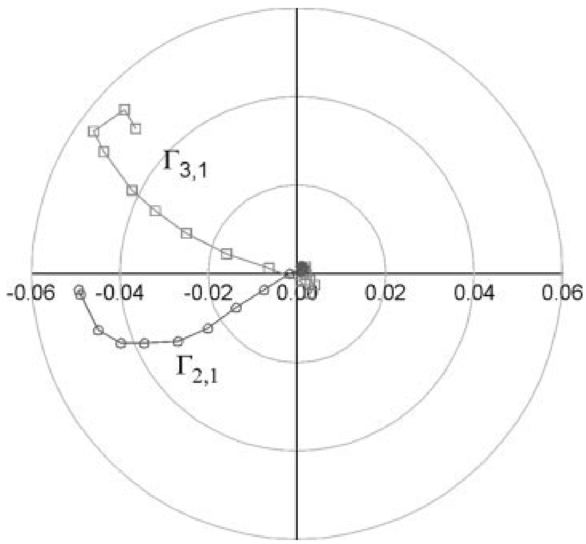


Figure 5. Conversion coefficients related to the up conversion for the upper, $\Gamma_{3,1}$, and lower, $\Gamma_{2,1}$, side bands.

different current contribution in the diode junction capacitance and in turn a rotation of the measured current phasors.

A second noticeable effect is related to the comparison between the upper side band and the lower side band up-conversion. Figure 4 shows the results of the injection of a signal at 650 MHz in terms of the measured conversion coefficients defined consistently to Eq. (1) as a function of the LO power. Increasing the LO level increases the reflection coefficients, as expected, with different amplitudes and phases, which is related to the nonlinear junction capacitance. Indeed, the LO signal determines conductance and capacitance waveforms whose Fourier series coefficients contribute to the CM features. In the case of an injected signal at the lowest IMP, the upper and lower side band up-converted signals refer to the same voltage phasor, which is multiplied by different capacitance Fourier series coefficients. This justifies the shape of the two conversion coefficients. Since the LO level is limited in the lower range, say, lower than +4 dBm, the phases of $\Gamma_{5,1}$ and $\Gamma_{4,1}$ are concentrated around the 90° axis, which reflects a dominant capacitance behavior. For higher LO levels, the time varying conductance becomes dominant and the resulting behavior is steered toward the 0° axis.

The subharmonic mixing performance is also compared with the up-conversion behavior associated with the LO fundamental, namely, LO $\pm$ IF. As expected from the basic conversion mechanisms, specifically, the device conductance nonlinearity of the third order, the mixing products associated with the fundamental are much lower than those associated with the second LO harmonic. This feature is illustrated in Figure 5, where the conversion coefficients $\Gamma_{3,1}$ and $\Gamma_{2,1}$ are compared.

Again, in this case the magnitude increases as the LO level increases, although they are lower by an

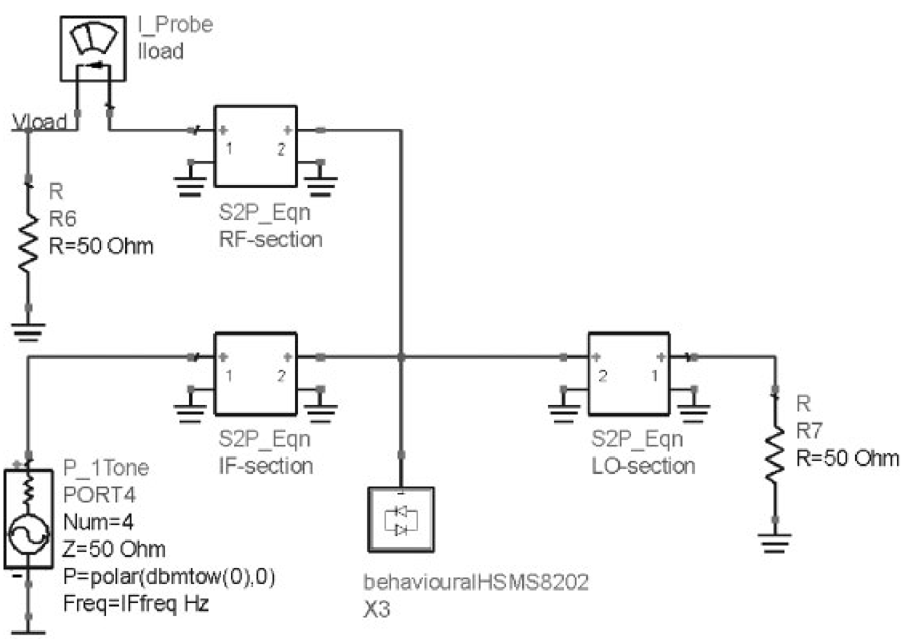


Figure 6. Schematic representation of the subharmonic mixer comprising the behavioral model of the Schottky-pair.

order of magnitude than the case of $\Gamma_{5,1}$ and $\Gamma_{4,1}$, confirming the attitude of this nonlinear cell for subharmonic mixing purposes.

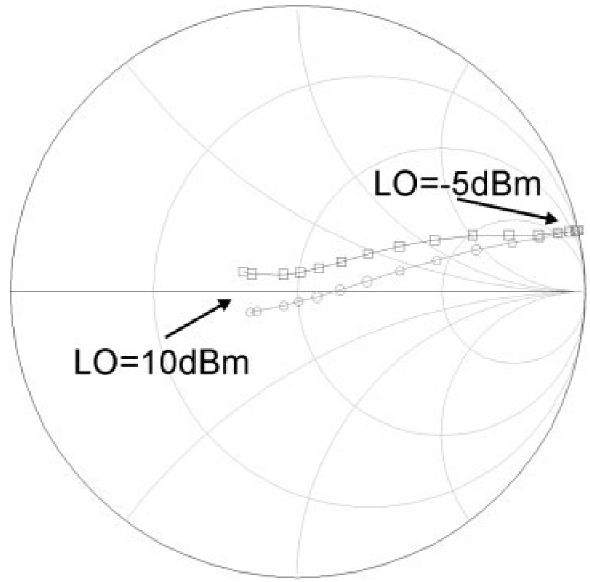
IV. DESIGN OF THE SUBHARMONIC MIXER

This section deals with the design of a microstrip subharmonic mixer based on the above described diode-pair with IF = 650 MHz, RF = 5.85 GHz, and LO = 2.6 GHz; the selected LO level is 8 dBm. The model represented in Figure 1 is introduced into the schematic representation of the subharmonic mixer as shown in Figure 6 for the purpose of the design of the linear parts of the circuits, namely, the triplexer for the LO-IF-RF signal combination, the RF and IF matching, and the matching of the LO and its harmonics. In particular, the latter is mandatory because it is a necessary condition that the CM is inserted into a circuit that preserves the LO terminations that were implemented in the measurement setup. This means that if the LO and its harmonics are terminated in a broadband 50Ω system, the same should be implemented in the circuit. In certain cases the latter may constitute an increase of complexity in the mixer design.

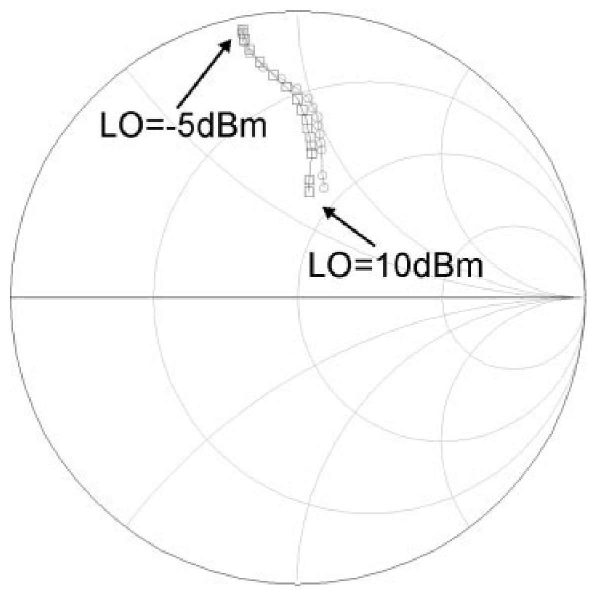
The design of the LO-, RF-, and IF-section is aimed at reaching the following ideal behavior [9]:

1. LO-section: $s_{11} = s_{22} = 0$ and $s_{12} = s_{21} = 1$ at LO frequency; $s_{22} = 0$ at LO harmonics; $s_{22} = 1$ all the IMPs;
2. RF-section: $s_{11} = 0$; s_{22} = *rf-optimum value* and $s_{12} = s_{21} = 1$ at RF frequency; $s_{22} = 1$ at LO fundamental and harmonics; $s_{22} = 1$ all other frequency;
3. IF-section: $s_{11} = 0$; s_{22} = *if-optimum value* and $s_{12} = s_{21} = 1$ at IF frequency; $s_{22} = 1$ at LO fundamental and LO harmonics; $s_{22} = 1$ all the IMPs;

The proposed method considers the *rf-* and *if-optimum value* arising from the linear analysis of the CM [8]. In that case, with the basic assumption that the unwanted IMPs are shorted, the optimum RF and IF terminations are calculated in a straightforward manner. Of course, the hypothesis of having the unwanted IMPs shorted is not easily verifiable in actual circuits, so for this reason a CAD analysis which considers the phase and amplitude combination between all the possible IMPs is required for an optimized circuit. In [9] the optimization of the performance is made by sweeping the phase of s_{11} of the RF-section at some selected IMPs and for each sweep the phase at the remaining frequency is kept equal to zero.



(a)



(b)

Figure 7. Γ_{opt}^{IF} (a) and Γ_{opt}^{RF} (b) as a function of the LO level calculated in the 50Ω system (squares) and at the diode section of the mixer at the end of the design procedure (circles).

The design of the linear sections in Figure 6 starts by considering the general features that any single block should have, i.e., introducing filters and resonators as is usually done in a conventional microwave design of a passive structure. Additionally, the search for optimum phases of the reflection coefficients as

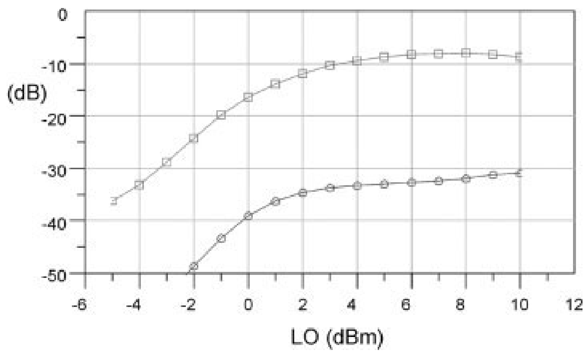


Figure 8. Conversion gain of the designed subharmonic mixer (squares) and response at the image frequency (circles).

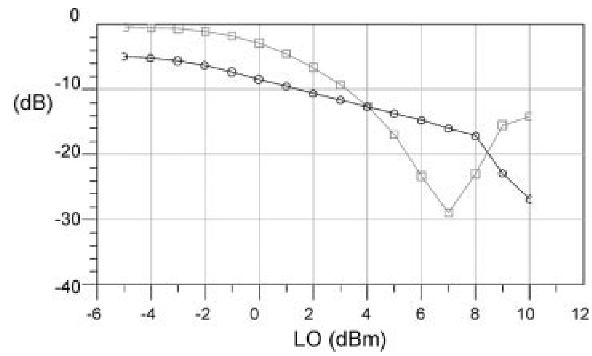


Figure 9. Reflection coefficients of the designed subharmonic mixer IF port (squares) and RF port (circles).

defined in the above list starts with considering the conjugate of the diode-pair reflection coefficients in a 50Ω system:

$$\Gamma_{opt}^{RF} = \frac{b(\omega_{RF})}{a(\omega_{RF})};\ \Gamma_{opt}^{IF} = \frac{b(\omega_{IF})}{a(\omega_{IF})}, \tag{2}$$

where the quantities in Eq. (2) are obtained by injection of a small signal at both the RF and IF, in presence of the LO, and evaluating the voltage wave variables a and b by the relations:

$$a = \frac{v = 50 \cdot i}{2};\ b = \frac{v - 50 \cdot i}{2}. \tag{3}$$

The design considers Eq. (2) as the starting point. During the iterations, new targets Γ_{opt}^{RF}, Γ_{opt}^{IF} are calculated. These values will be different from the previous iterations because the broadband frequency behavior of the passive networks modifies the exhibited impedance and in turn modify the targeted values. This mechanism is due to the CM, which combines all the IMPs voltages to give the actual current at the frequency of interest. The convergence of the design process is ensured by its linear nature and by the small deviations from the initial guesses.

Figure 7a,b shows the Γ_{opt}^{IF} and Γ_{opt}^{RF} , respectively, in the 50Ω system and calculated at the end of the design process as a function of the LO level.

Figure 8 shows the up-conversion gain compared with the image response. The rejection of the image frequency is due to the selectivity of the sub-blocks, which are based on two-section band pass filters and a resonator which allows the simulated rejection. Figure 9 plots the reflection coefficients at the RF and IF ports as a function of the LO level. It is observed that both ports show a minimum for values slightly different from the specified. An improved matching condition is attainable only at the expense of complexity and does not appear justifiable. The results of the design procedure in terms of the microstrip layout is reported in Figure 10, from which we can easily observe the sub-blocks and their respective rules.

V. CONCLUSION

This article has dealt with the possibility of using the conversion matrix of an antiparallel Schottky diode pair as a behavioral model for the purpose of analyzing and designing mixers. The analysis is simplified by the presence of a behavioral model extracted from measurements instead of conventional equivalent circuit models. The simulation involves only linear components and frequency-domain defined device parts and uses a single tone harmonic balance engine to produce the desired intermodulation products. The

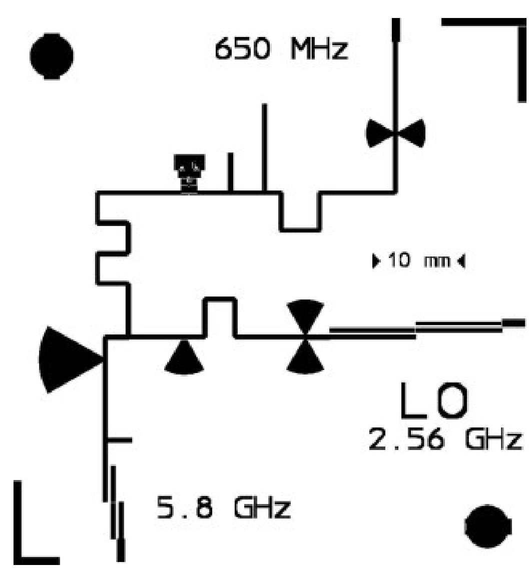


Figure 10. Layout of the designed subharmonic mixer.

proposed design procedure appears straightforward, although it does require the additional constraints of an LO matched at the nonlinear device terminations. Presently the approach is suitable only for the selected frequency plan, although an extension to a bandwidth around the IF and RF appears feasible. The article has also dealt with an example of application of the proposed approach to the design of subharmonic mixer.

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BIOGRAPHIES

Alessandro Cidronali was born in Florence, Italy, in 1965. He received the Laurea and Ph.D. degrees in electronic engineering from the University of Florence, Florence, Italy, in 1992 and 1998, respectively. In 1993, he joined the Department of Electronics Engineering, University of Florence, where he became an Assistant Professor in 1999. During his academic career, he has been a Lecturer in courses on "Applied Electronics" and "Solid-State Electronics" and currently teaches "Microwave Electronics." His research activities cover the study of analysis and synthesis methods for nonlinear microwave circuits, the design of broadband MMICs, and the development of CAD and numerical modeling for microwave devices and circuits. He is a Guest Researcher at the National Institute of Standards and Technology (NIST), Electromagnetics Division, Non-Linear Device Characterization Group. In the frame of the EU Network TARGET – "Top Amplifier Research Groups in a European Team," supported by the information Society Technologies Program, IST-1-507893-NOE, he serves as Workpackage Leader for the "Transmitters modeling\architectures for wireless broadband access" workpackages. Dr. Cidronali is currently an associate editor for the IEEE Transaction on Microwave Theory and Technique.

Giovanni Loglio was born in Florence, Italy, in 1978. He received a Master's Degree from Universita' degli Studi of Florence, Italy, in 2003. He is a Ph.D. student at Universita' degli Studi of Florence, Italy, and at the University of Colorado at Boulder jointly. He is a Guest Researcher at NIST, Electromagnetics Division, Non-Linear Device Characterization Group. His main research interests are concerned with nonlinear device and mixer circuit characterization for applications in the microwave range.

Jeffrey A. Jargon received the B.S., M.S., and Ph.D. degrees in electrical engineering from the University of Colorado at Boulder in 1990, 1996, and 2003, respectively. He has been with the Electromagnetics Division of the National Institute of Standards and Technology (NIST), Boulder, CO, since 1990. His current research interests include calibration techniques for nonlinear vector network analyzers and artificial neural network modeling of passive and active devices. Jeffrey has authored or co-authored over 40 technical publications and conference presentations, receiving three awards for best paper, an URSI Young Scientist Award, and a Department of Commerce Silver Medal Award. He is a member of Tau Beta Pi and Etta Kappa Nu, is a Senior Member of IEEE, and is a registered Professional Engineer in Colorado. Jeffrey also currently serves as the Chair of the IEEE MTT-S Measurements Technical Committee and is an Adjoint Professor at the University of Colorado at Boulder.

Gianfranco Manes was born in Florence, Italy, in 1944. He became Associate Professor in 1980 and Full Professor in 1985. Dr. Manes contributed, from an early stage, in the field of Surface-Acoustic-Wave (SAW) technology for RADAR signal processing and Electronics Countermeasure applications. Major contributions were in introducing novel FIR synthesis techniques, fast analogue spectrum analysis configurations, and Frequency Hopping waveform synthesis. Since the early 1980s Dr. Manes has been active in the field of microwave modeling and design. In the early 1990s he founded and is currently leading the Microelectronics Lab of the University of Florence, committed to research in the field of Microwave Devices. In 1982 he committed to build up a facility for the design and production of SAW and MIC/MMIC devices, as a subsidiary of a Florence' Radar Company, SMA Spa. In 1984 the facility became a stand-alone, privately owned microwave company, Micrel SpA, operating in the field of Defence Electronics and Space Communications. In the period 1996–2000 he was involved in IV framework projects, in the field of Information Technology applied to Cultural Heritage, and was invited to orientation meetings and advisory panels for the Commission. Present research interest is in the field of RITD devices for microwave applications in a scientific collaboration with the group at the Physical Science Research Labs, Motorola Corp., Tempe, Arizona. Dr. Manes was founder and is presently President of MIDRA, a research consortium between the University of Florence and Motorola. He is a member of the Board of Italian Electronics Society and Director of the Italian PhD School in Electronics. In November 2000 he was appointed Deputy Rector for the Information System of the University of Florence. Dr. Manes is currently a member of the International Microwave Symposium Technical Program Committee.