

Short Papers

Accurate Analysis and Design of Millimeter Wave Mixers

Franco De Flaviis, Tullio Rozzi, Franco Moglie, Andrea Sgreccia, and Anselmo Panzeri

Abstract—In recent years, various techniques have been developed for the study of microwave mixers. The present contribution stems from the method introduced in [1] for the nonlinear analysis, but differs in the linear part. In particular, we emphasize the circuit characterization of the mixer harmonics, removing the usual approximation of matched terminations and employing models that effectively represent network characteristics in the microwave and millimeter wave domain. This goal is achieved through the accurate knowledge of the various components occurring in the mixer and the use of accurate equivalent circuits. In order to validate the model, two subharmonic mixers operating at *K*-band (18–26 GHz) were realized. Reasonable agreement is achieved between theoretical and experimental results.

I. INTRODUCTION

In the course of recent years, many authors have treated the study of microwave mixers and developed techniques for their design, for instance [2]–[6] to quote just a few. Moreover, an excellent and up to date list of reference is provided in [1].

Our work utilizes the nonlinear analysis method introduced in [1] that appears to be the most effective so far in obtaining convergence. We differ, however, in the linear part of the analysis, in that we remove the usual approximation of matched terminations for the mixing products. A fundamental problem in millimeter wave mixer research consists in determining accurately the behaviour of the embedding circuits over large frequency ranges. We have addressed this problem by using the Hewlett-Packard “MDS” microwave simulation package, complementing it with our results [7] and extending its accuracy and frequency range. This allowed accurate modelling of the behavior of the linear net over a very large frequency range. We designed, built and tested two subharmonic mixers (see Figs. 1 and 2) operating in *K*-band using planar technology in thin film, one in microstrip, one in slotline. Reasonable agreement between predicted and experimental conversion losses shows our model to be both efficient and accurate and to constitute an effective basis for design.

II. DESIGN DATA

The harmonic mixing has been used primarily at the higher millimeter wave frequencies where reliable, stable LO generators are difficult to realize and at same time expensive. The conversion loss obtained by harmonic mixing has been typically 3 to 5 dB greater than that which could be obtained by fundamental mixing at the same signal frequency [8], [9].

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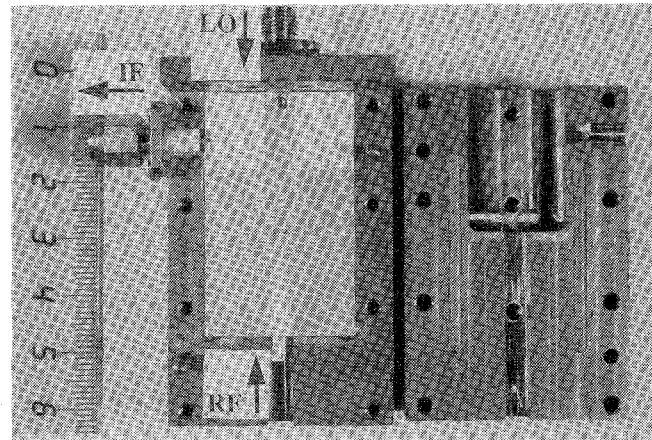


Fig. 1. Photograph of subharmonic mixer 1 operating as down-converter with $f_{IF} = 70$ MHz, $f_{LO} = 11.5$ GHz and $f_{RF} = 23.07$ GHz.

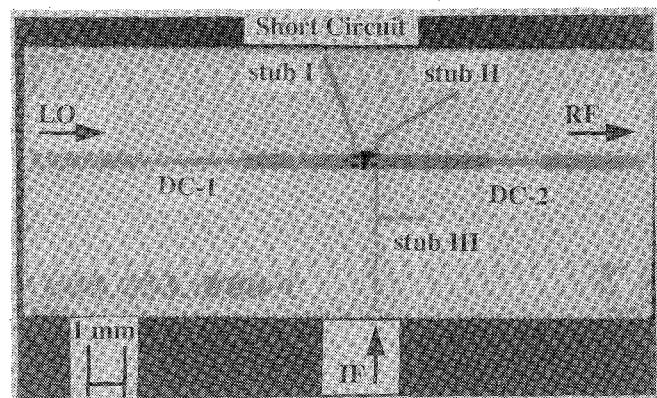


Fig. 2. Photograph of subharmonic mixer 2 operating as up-converter with $f_{IF} = 70$ MHz, $f_{LO} = 11.5$ GHz and $f_{RF} = 23.07$ GHz.

We have realized two subharmonic mixers in thin film planar technology using 0.38 mm thick alumina substrates and Au metallization. The first utilizes various planar guides, the second just microstrip. For more details see Figs. 1 and 2.

The diodes used were beam-lead Schottky diodes HP HSCH-5340. The characteristic data of this are:

$$I_s = 0.6E - 9 A$$

$$C_p = 0.14 \text{ pF}$$

$$C_{jo} = 0.19 \text{ pF}$$

$$L_s = 0.1 \text{ nH}$$

$$\gamma = 0.5$$

$$R_s = 5 \Omega$$

$$\eta = 1.08$$

which are taken from measurements and data sheets.

A. Subharmonic Mixer 1

Fig. 1 shows the configuration of balanced subharmonic mixer 1, operating as a down-converter. This was realized by means of rectangular waveguide, cut along its broad side, wherein is placed an alumina plate (0.38 mm thick). The LO signal at 11.5 GHz, propagates along suspended microstrip, the RF signal at 23.07 GHz along the waveguide; the IF signal at 70 MHz is extracted by means of a line in suspended microstrip. The equivalent circuit was implemented on the "MDS" simulation model of the slotline to guide transition, containing a cascade of slotlines of varying characteristic impedance; all ports are closed on $Z_0 = 50 \Omega$ loads, typical of microwave circuits.

B. Subharmonic Mixer 2

Fig. 2 depicts the configuration of the subharmonic mixer operating as up-converter. This is realized using solely microstrip lines with gold metallization (3 μm) grown on a 0.38 mm thick alumina substrate. The key element of the mixer's operation lies in that the RF frequency is just double that of the LO signal, i.e. $f_{\text{RF}} = 2f_{\text{LO}}$ [10].

Consequently, a stub of length $\lambda_0/4$ having infinite impedance at a given section for the LO signal becomes a short circuit for the RF signal and viceversa [11]. In order to prevent the IF signal from exiting from the RF port (IF/RF isolation), we have utilized a DC block (DC-2), centered at 23.07 GHz that provides further LO/RF isolation, in addition to that already provided by stubs I and II. Isolation between RF and IF is obtained by means of stub III, that produces an open circuit for the RF signal at the diodes section and prevents it from propagating through to the IF.

In the equivalent circuit implemented on the "MDS," all ports are closed on $Z_0 = 50 \Omega$ loads and the embedding impedance seen at the diode terminals is given by

$$Z_{\text{emb}} = 2Z_0 \frac{1 - S_{11}S_{22}}{(1 - S_{11})(1 - S_{22})} \quad (1)$$

where S_{11} and S_{22} are respectively the reflection coefficients at port-1 and port-2 when the diodes are connected.

Fig. 3 shows the results of the simulations on "MDS," obtained for the subharmonic mixer 2 over the range from DC to 50 GHz.

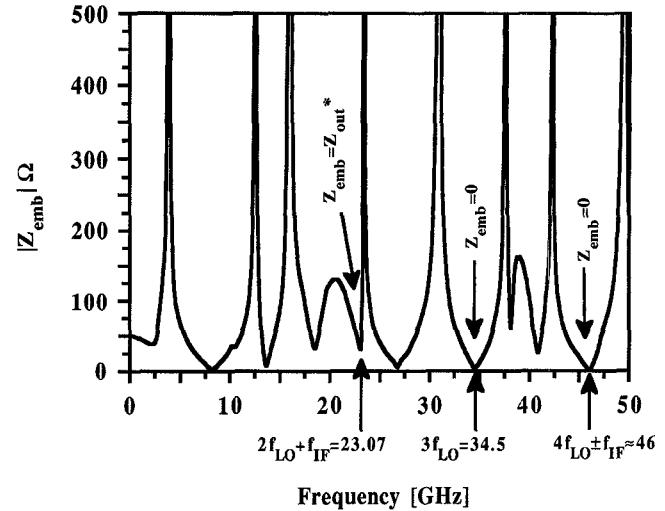


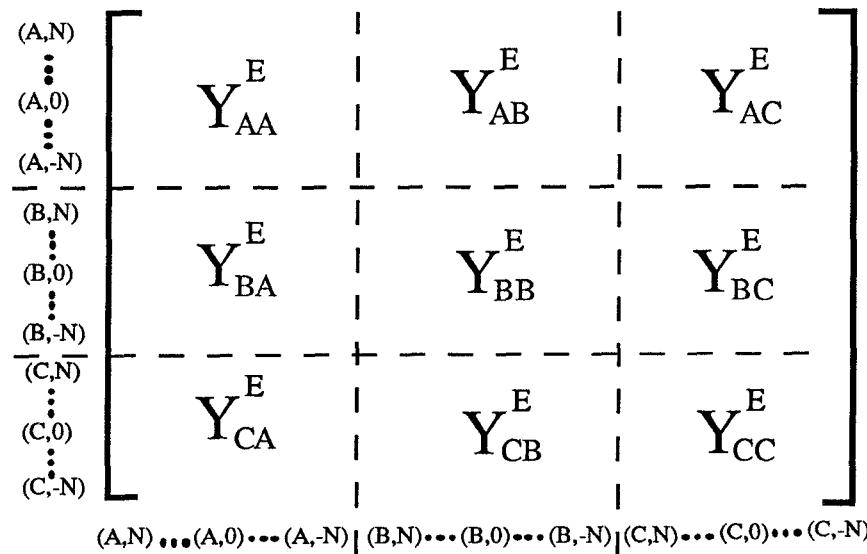
Fig. 3. Simulation on "MDS" of subharmonic mixer 2 over range from 0 to 50 GHz. We observe that the impedance that the diodes see at his terminals is approximately 0Ω at $3f_{\text{LO}}$ and at $4f_{\text{LO}} \pm f_{\text{IF}}$.

III. POWER FLOW IN THE SUBHARMONIC MIXER

We have realized two subharmonic mixers, that differ from the usual in that the pump frequency (LO) is just half of pump frequency of usual mixer [10]. It is important to know which frequency must be well matched in order to obtain good performance in terms of conversion losses. The knowledge of the conversion loss for a subharmonic mixer is obtained from a knowledge of his admittance matrix [12].

A double-index port numbering system is used [12]. The embedding network is characterized by the admittance matrix Y^E , a typical element of which is: $Y_{(A,m)(B,n)}^E$; row and column numbering of Y^E are reported in (2) at the bottom of the page; $Y_{i,j}^E$ is a submatrix of Y^E that represents the coupling between the band 'i' ($\omega_i = i\omega_{\text{LO}} + \omega_{\text{IF}}$) and the band 'j'. If the network is linear and time-invariant there is no internal coupling between ports at different sideband frequencies, hence the only nonzero elements of Y^E are those for $i = j$.

If the conversion matrix of the two pumped diodes [12] are denoted Y^A and Y^B , the parallel connection of the two diodes with the



embedding network results in an overall mixer admittance matrix [12]

$$Y^M = \begin{bmatrix} Y^A + Y_{AA}^E & Y_{AB}^E & Y_{AC}^E \\ Y_{BA}^E & Y^B + Y_{BB}^E & Y_{BC}^E \\ Y_{CA}^E & Y_{CB}^E & Y_{CC}^E \end{bmatrix} \quad (3)$$

In order to obtain the IF output impedance of the mixer, the IF load admittance is set to zero, so that the impedance seen looking into port $(c, 0)$ is simply the output impedance.

Let $Y^{M_{c/o}}$ denote the admittance matrix of the mixer with zero load admittance (open circuit), and define

$$Z^{M_{c/o}} \triangleq (Y^{M_{c/o}})^{-1} \quad (4)$$

the output impedance of the mixer is given by element $(c, 0)(c, 0)$ of $Z^{M_{c/o}}$ i.e.

$$Z_{\text{OUT}} = Z_{(c,0)(c,0)}^{M_{c/o}} \quad (5)$$

for optimum conversion loss the output of the mixer should be conjugate-matched. Inverting the mixer admittance matrix gives the mixer impedance matrix

$$Z^M \triangleq (Y^M)^{-1} \quad (6)$$

if a small signal $\delta I_{(c,k)}$ at sideband ω_k is applied at port (c, k) of the mixer, the output response at port $(c, 0)$ is

$$\delta V_{(c,0)} = Z_{(c,0)(c,0)}^M \delta I_{(c,k)} \quad (7)$$

the power delivered from the load is

$$R_{\text{delivered}} = |\delta V_{(c,0)}|^2 \text{Re}[Y_0]^2 \quad (8)$$

and the power available from the source is

$$P_{\text{available}} = \frac{|\delta I_{(c,k)}|^2}{4 \text{Re}[Y_k]} \quad (9)$$

hence the conversion loss from sideband ω_k to sideband ω_o is given by

$$L_{o,k} = \frac{1}{4|Z_{(c,0)(c,k)}^M|^2 \text{Re}[Y_k] \text{Re}[Y_o]} \quad (10)$$

and consequently the conversion loss from sideband ω_j to sideband ω_i is

$$L_{i,j} = \frac{1}{4|Z_{(c,i)(c,j)}^M|^2 \text{Re}[Y_j] \text{Re}[Y_i]} \quad (11)$$

Through relation (11) we can calculate the power flow in the subharmonic mixer at each frequency. A 11.5 GHz LO input and a 0.07 GHz low-level signal were impressed at the slot line input. We observe the absence of fundamental mixing products ($f_{\text{LO}} - f_{\text{IF}}$) and ($f_{\text{LO}} + f_{\text{IF}}$) and of all terms $n f_{\text{LO}} \pm m f_{\text{IF}}$ with $m+n$ even. The main power flow takes place in the two sideband USB ($2f_{\text{LO}} + f_{\text{IF}}$) and LSB ($2f_{\text{LO}} - f_{\text{IF}}$), but it is important observe how some power also flows in the bands ($4f_{\text{LO}} - f_{\text{IF}}$) and ($4f_{\text{LO}} + f_{\text{IF}}$) and also the output at $3f_{\text{LO}}$ is much greater than that at $2f_{\text{LO}}$. In conclusion, it is important remark that all terms at frequency $f_k = |k f_{\text{LO}} + f_{\text{IF}}|$, where k is even, must be well matched for good mixer design. In our case we have matched the USB with his conjugate load and short circuited the other sideband. The LSB is difficult to match in different manner than the USB, as they are very near (only 140 MHz of difference), but we can remove this term by means of a simple SSB configuration [6].

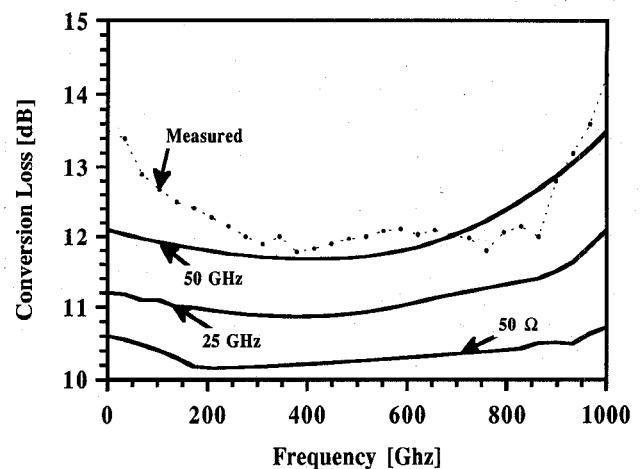


Fig. 4. Conversion loss for subharmonic mixer 1 (theory and experiment).

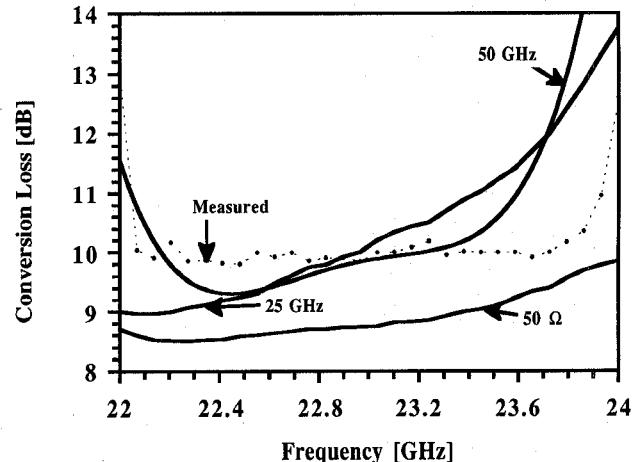


Fig. 5. Conversion loss for subharmonic mixer 2 (theory and experiment).

IV. RESULTS

In Figs. 4 and 5 are reported the experimental and theoretical curves of conversion loss vs. frequency for the two mixers under study.

The curves labelled by "50 GHz" were obtained using our CAD analysis, where we provided as input a description of the microwave network valid up to 50 GHz. This is given as frequency-dependent impedance seen at the diode terminals, by considering all the ports of the network (source and load ports) as closed by constants 50Ω loads. Such description of the network is obtained, as previously stated with the help of the "MDS" program corrected by means of the results of [13]. In order to highlight the degree of accuracy introduced by this kind of analysis, we have reported the theoretical curve obtained by computing the conversion loss in the standard manner, that is, by neglecting the frequency dependence of the network impedance and considering the latter as a constant 50Ω load; this curve is labelled "50 Ω". The estimate of conversion loss obtained by our model is, in some cases as much as 2 dB more accurate, corresponding to a 20% improvement.

The curve labelled respectively "25 GHz" and "50 GHz" shows in which measure knowledge of the network up to a given frequency does influence the estimate of conversion loss. In our case, we did not take the model beyond that of the fourth harmonic of LO, since the impedance of the first four harmonics are of greater importance anyway, the "MDS" models can no longer be relied upon beyond these frequencies and moreover higher frequencies are certainly short

circuited by the diode capacitances that at frequencies above 50 GHz present a reactance less than 10Ω .

Fig. 5 also show that for mixer 2 the bandwidth it is flat over a band of nearly 2 GHz.

V. CONCLUSIONS

We have developed a sophisticated model for millimeter wave mixers, featuring a new, particularly accurate linear analysis portion that takes into account the terminations of all relevant mixing products over a wide band. The good degree of agreement between experimental and theoretical data confirms the effectiveness of the model.

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A Design Method for Lumped Broad-Band MMIC Matching Networks with Semiuniform Frequency-Dependent Losses

Quinan Lin, Hamid Aghvami, and Ian Robertson

Abstract—A new design method for broad-band MMIC matching networks, which consist of lumped inductors and capacitors, is presented in this paper. Based on the fictitious transformation between the lossy network and the lossless one [6] the transducer power gain (TPG) of the lossy network is calculated using impedance matrix and transmission parameter matrix methods. Then, on the basis of the result of TPG optimization, the matching network can be synthesized. Since only the complexity of the matching network needs to be specified, so the advantage of the "real frequency technique" [5] is retained. In this design procedure the frequency-dependent losses of lumped elements can be considered. Thus the actual gain response tends to coincide with the desired performance more than when the losses of elements are neglected. An example is given to show the application of the new method to broad-band GaAs FET amplifier design.

I. INTRODUCTION

For the synthesis of matching networks for broad-band GaAs MESFET amplifiers, the elements of the network have generally been considered as lossless in the previously reported techniques. In order to occupy less GaAs chip area, the matching networks for MMICs are usually constructed by lumped or semi-lumped reactance elements. Because these elements have small size and operate in the microwave frequency range, neglecting the losses would cause rather evident errors in design [1]. Therefore it is necessary in practice to improve the synthesis procedure of the matching networks. Some approaches to the synthesis of broad-band matching networks with losses have been reported in the past few years. Based on Andersen's transform [2], Youla's analytical theory of broad-band matching can be extended to the lossy network of the semiuniform type (i.e. all inductors have one quality factor Q ; all capacitors have another Q [3]). In the design of microwave broad-band amplifiers, if the unilateral model of the GaAs FET is used and the losses of the circuit elements are assumed frequency independent, the arbitrary non-uniform lossy network can be synthesized [4]. By use of the fictitious transformation between the lossy network and the lossless one and using the "real frequency technique" [5] the lumped matching network with semiuniform frequency-dependent losses can be designed [6] and a lossy commensurate line network can be synthesized [7].

In this paper, we present a new synthesis method for the lumped broad-band matching network with semiuniform frequency-dependent losses. According to the fictitious transformation, the impedance parameters of the lossy network can be obtained from the lossless one. Then the transducer power gain of the lossy network is calculated using transmission parameter matrices and optimized using the least-square method. On the basis of the result of TPG optimization, the synthesis of the matching network can be implemented in a fictitious transformation domain. It is more direct

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