Multitone Characterization and Design of FET Resistive Mixers Based on Combined Active Source–Pull/Load–Pull Techniques

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***Abstract*— A dual six-port-based measurement setup was developed to synthesize five source and load impedances simultaneously. The setup can perform nonlinear measurements with multifrequency excitation. Active source–pull/load–pull measurements obtained for an NE-9001 transistor operated in a C-band field-effect transistor (FET) resistive mixer mode allow one to optimize the linearity of the mixer while maintaining a typical conversion loss of approximately 7 dB. Two-tone verification at 3.9000 and 3.9005 GHz showed that the level of in-band third-order intermodulation products could be reduced to -50 dBc, with a well-chosen output intermediate frequency (IF) load impedance and sufficient local oscillator (LO) power. The measured performance of the realized mixer is in good agreement with that predicted at the transistor characterization step of the design.**

***Index Terms*— Characterization, intermodulating, load–pull, nonlinear, resistive mixer, six-port.**

I. Introduction

IN-BAND intermodulation (IBIM) performance is an important characteristic to consider in the design of microwave/millimeter-wave front-end mixers. In a multicarrier or digital communications system, IBIM products generated by the mixer can seriously deteriorate the main intermediate frequency (IF) signal. The field-effect transistor (FET) resistive mixer was proposed by Maas in 1987 [1] as an answer to this problem. Since then, several reports on the FET resistive mixer have been published, showing good performance of this type of FET mixer in terms of conversion loss, higher operating frequency, and low level of intermodulation (IM) products [2]–[5]. Most of the works were based on the use of a nonlinear model with computer-aided design (CAD) software and only few were based on experimentation. For the time being, it is still relatively difficult to perform multitone characterization of nonlinear devices using conventional instruments. Various methods and nonlinear measurement setups aimed at the design of solid-state power amplifiers (SSPA's) have been reported [6]–[9]. Based mainly on the load–pull principle, these techniques use various instruments such as power meters, a spectrum analyzer, a vector network analyzer (VNA) [6] or a six-port junction [9], computerized tuners [8], etc., to obtain large-signal constant-output power contours for the single-tone test and third-order intermodulation contours for the two-tone test. These measurement test sets are often cumbersome, generally require tedious labor, and sometime cannot access needed information. Furthermore, applying such SSPA measurement methods or setups to the characterization of mixers demands many hardware modifications, which can easily increase the complexity of the test set. These technical constraints and inconveniences limit the accuracy and predictability of microwave single-ended FET mixer design.

This paper presents an experimental approach to the design of very wide-dynamic-range (linear) FET resistive mixers. Using two six-port reflectometers, an active source–pull measurement (ASPM) and load–pull measurement system was developed to study constant power contours as functions of the local oscillator (LO) and radio-frequency (RF) source impedances and the LO, 2LO, and IF load impedances. Active source–pull measurement allows one to report on important effects of the LO source impedance on the undesirable 2LO signal and on the spurious responses of the FET resistive mixer. Both single- and two-tone characterization were investigated. This paper also discusses some problems encountered during the assembly of the test set and the hardware solutions finally adopted.

II. Description of the Mutifrequency Setup

In this section, the active source–pull technique is first described in order to establish a bases for the following discussion on the active source–pull/load–pull multifrequency test set.

A. Background for the ASPM

The source–pull measurement is aimed at evaluating linear/nonlinear device performance as a function of source impedance. Commonly used passive source–pull techniques [7], [11], which are based on the use of a tuner, are unable to independently control the variation of the source impedance over some frequency band. It is also difficult to present a quasi-unitary source reflection coefficient when the loss in the setup is not negligible. To overcome these limitations, the passive source–pull technique using a reverse six-port reflectometer [12] was improved. The idea is very simple. Fig. 1(a) shows

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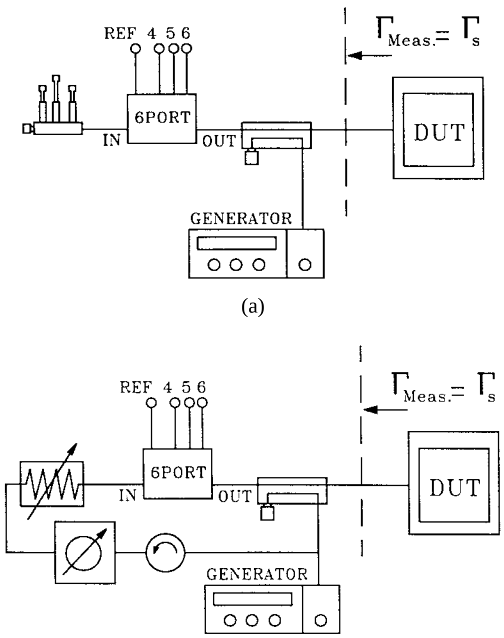


Fig. 1. Six-port-based active source–pull techniques. (a) Passive source-pulling using stub tuner. (b) Active source–pulling.

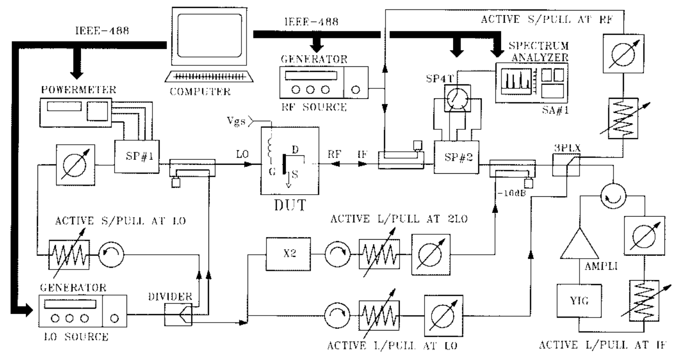


Fig. 2. Multifrequency test set.

the setup for a passive source–pull: the input signal is injected into the device-under-test (DUT) via a directional coupler, and the source impedance is set with a stub tuner. The six-port in such an arrangement measures the source impedance seen at the reference plane. Since the source mismatch is due to the wave reflected by the stub tuner, the source impedance can be synthesized in an active manner, similar to the active load–pull technique. In fact, by injecting a part of the input signal well controlled in amplitude and phase [see Fig. 1(b)], the reflected wave can be synthesized. The total loss in the setup is easily compensated for by increasing the amplitude of the synthesized reflected signal. Thus, a truly unitary source reflection coefficient can be produced at the DUT reference plane.

This suggested ASPM technique was verified by comparing the measured available gains of a linear amplifier with the theoretical values [12]. It was proven that a truly unitary source reflection coefficient can be attained. The proposed ASPM technique provides an accuracy of $\pm 0.007\angle\pm 0.5^\circ$ for reflection coefficient measurements and ± 0.1 dB for available power evaluations, even in the critical region where the available gain of the test device is sensitive to matching conditions. Using separate active source–pull loops and a power combiner, it is also possible to individually adjust the source impedance at each input frequency.

B. Multifrequency Measurement Test Set

The block diagram in Fig. 2 shows the main components of the setup. It is a dual six-port network analyzer modified in order to simultaneously perform an ASPM and load–pull measurement on an FET resistive mixer. The DUT, an NE 9001 GaAs FET, receives the LO and RF signals at the gate and drain, respectively. The IF signal is extracted from the drain. Notice that since the drain is not biased, the transistor is used as a time-varying resistance controlled by the LO signal.

The six-port junction SP#1 and SP#2 are designed to have low insertion loss, and can operate from 2 to 5.6 GHz and 0.5 to 9 GHz, respectively. The power at four detection side arms of the SP#1 is measured by a four-channel power meter. A computer-controlled SP4T switch allows one to successively measure the powers at the four detection ports of the SP#2. Since many frequencies are simultaneously present at the output, there are two possible ways to take power readings at the ports of the SP#2: either by using a power probe, along with a computer-controlled yttrium iron garnet (YIG) filter, or by using a spectrum analyzer controlled via an IEEE-488 bus and calibrated in advance for precise power measurements.

The first solution has the advantage of accuracy and speed of measurements. However, the following three main problems in mixer characterization may favor the use of a spectrum analyzer.

1) The stability of the computer-controlled YIG filter versus the variations in the temperature and polarized current. Good control of these parameters is always possible, but intensive labor and (costly) setup complexity are unavoidable.
2) In a two-tone test, the third- and fifth-order IBIM products are present inside the 30-MHz passband of the YIG filter.
3) A very weak signal at the detection ports of the SP#2 six-port. In fact, when characterizing low-power nonlinear devices such as FET resistive mixers, the power which needs to be detected by the probe can be below -50 dBm and be submerged in the noise of the YIG filter and power sensor. In such a case, the dynamic range of detection (i.e., the allowed variation of power at each detection port of the SP#2) might be as low as 15 dB, which is often insufficient for practical needs.

A high-quality spectrum analyzer (SA#1) can circumvent these problems. A very weak signal becomes measurable when a high-resolution bandwidth is used. The dynamic range of SP#2 can be increased to 60 dB, which is sufficient to carry out measurements over the frequencies of the intermodulation products of interest. In addition, the presence of a spectrum analyzer in the SP#2 allows one to observe any higher order products of the LO and RF signal (spurious responses). An appropriate setting of the spectrum analyzer could give both reasonable accuracy and speed of measurement.[1]

A conventional six-to-four-port reduction technique [13] is employed to calibrate the SP#1 and SP#2 junctions. The thru-reflection line (TRL) method is used to de-embed the measurements at the input-output (I/O) planes of the transistor, which is mounted on a microstripline test fixture. Since only the LO signal is injected into the gate and several signals are present at the drain of the transistor, the two six-port junctions are calibrated as follows:

- SP#1 at 2.225 GHz (LO frequency) and SP#2 at 2.225 GHz;
- 3.600–3.900–4.200 GHz (IF frequencies);
- 4.450 GHz (2LO frequency);
- 5.825–6.125–6.425 GHz (RF frequencies).

The repeatability and the accuracy of the six-port SP#1 are within $\pm 0.005\angle\pm 0.5°$ for reflection coefficient measurement. The SP#2 overall measurement accuracy is around $\pm 0.015\angle\pm 3°$, which is sufficient for the matching circuit-design purposes.

The six-port reflectometer SP#1 performs LO active source–pulling at the transistor gate. The six-port junction SP#2 simultaneously holds ASPM at the input RF frequency and active load–pull measurements at the output LO, 2LO, and IF frequencies, both at the drain of the transistor. The power available from the source and the power absorbed by the load can be evaluated by SP#1 and SP#2, respectively, according to the following formulas [11], [14]:

$$P_{\mathrm{AVS}}^{\mathrm{SP\#1}} = \frac{1-|\Gamma_s|^2}{|1-S_{11}\Gamma_s|^2}|S_{21}|\frac{1}{1-|\Gamma_s|^2}P_{\mathrm{source}} \tag{1}$$

and

$$P_{\mathrm{ABS}}^{\mathrm{SP\#1}} = \frac{kP_{\mathrm{ref\text{-}port}}^{\mathrm{SP\#2}}}{|1+c\Gamma_L|^2} \tag{2}$$

where S_{ij} are the S-parameters of the two-port network defined between the generator and the measurement plane of SP#1. Γ_s and Γ_L are the reflection coefficients measured by SP#1 and SP#2, respectively, k is a real power-calibration constant, and c is the complex-error box constant.

A triplexer separates the LO, RF, and IF frequencies into three branches, while the 2LO active load–pull branch is realized by injecting the 2LO signal via a wide-band 16-dB directional coupler. Notice that the IF active load–pulling is performed by a closed loop in order to insure the frequency coherence of the IF reinjected signal. The insertion of a computer-controlled YIG filter in this closed loop eliminates any risk of oscillation. A quasi-unitary load can be attained at the DUT plane by increasing the synthesized signals to compensate for the combined insertion loss of the triplexer, six-port junction, 16-dB directional couplers, bias tee, and test fixture.

Two 2–18-GHz synthesized signal generators, along with additional power amplifiers, yield +15- and −4-dBm available powers for the LO and RF signals, respectively. The above maximum power levels are measured when the source impedance is set to 50 Ω in order to be matched to the 50-Ω power probe. It was also verified that the measured reflection coefficient of a fixed impedance remains quasi-constant when the source (LO and RF) powers vary from −35 dBm to their maximum available values. These variations define the dynamic range of the setup. The whole system is controlled by programs written in HP-Veetest, installed on an 80 486-based computer. Instruments are activated turn-by-turn during measurement via an HPIB controller. The YIG filter and the SP4T are controlled directly by a digital-to-analog 16-channel converter card. Fully automatic measurements can be done by replacing the manual phase shifters and variable attenuators with electronic vector modulators (EVM's).

[1] B. Peterson, "Spectrum Analysis Basics," Hewlett-Packard Appl. Note 150.

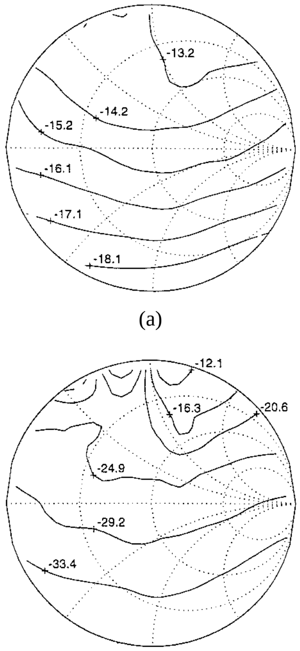
III. Experimental Results

The main objective of this paper was the characterization of the IBIM performance of a C-band FET resistive mixer as a function of the termination impedances. The more important parameters are optimized first and the less significant ones last, in all cases with single-tone RF excitation. Hence, the biasing point and input LO and RF power levels arrive first, followed by the LO and RF source impedance, respectively, and finally by the load impedances. These load impedances, in turn, are in order of importance: the load at the IF, then after the LO, and finally, at the second harmonic 2LO frequencies of the output signal. Once the effects of the termination impedances are determined, a two-tone RF excitation can be applied in order to study the IBIM performance of the transistor. These experimental steps are presented in the following sections.

A. Single-Tone Measurements

1) Bias Point and Input Power Levels: The FET resistive mixer is operated with a nonbiased drain ($V_{\mathrm{ds}} = 0$). In an ideal case, the gate bias voltage is chosen as the mean value between the gate pinchoff voltage and the gate built-in voltage. The LO signal is applied to the gate so that the drain-to-source resistance R_{ds} varies periodically in time between the lowest and highest resistance states. The unwanted intermodulation product is then minimized. The main drawback, however, is a significant conversion loss of typically −35 dB or more. In practice, the gate bias voltage is often chosen near or even below the pinchoff voltage. Thus, R_{ds} is retained for a longer time at its highest resistance state, and this additional effect (i.e., change in shape of $R_{\mathrm{ds}}(t)$ waveform) helps to reduce the conversion loss. In this paper, the NE 9001 transistor is used with a gate bias voltage of −3 V, which is below its pinchoff voltage of −2.6 V (at $V_{\mathrm{ds}} = 2$ V). The input power levels are fixed at $P_s^{\mathrm{LO}} = +6$ dBm and $P_s^{\mathrm{RF}} = -7.5$ dBm, which represent a practical level for the LO source and a level of the RF signal suitable for experimental characterization.

2) Optimization of the LO Source Impedance: Active source– pull measurement is performed at the gate of the transistor while passive loading conditions are set at the drain. The drain of the transistor is directly connected to the SP#2. Hence, the impedances of the measurement port of the SP#2 correspond to the predetermined values Γ_L^{LO}



(a)

(b)

Fig. 3. Optimization of the LO source impedance. (a) $P_L^{\rm IF}$ (3.900) in dBm versus $\Gamma_s^{\rm LO}$ (2.225). (b) $P_L^{\rm 2LO}$ (4.450) in dBm versus $\Gamma_s^{\rm LO}$ (2.225).

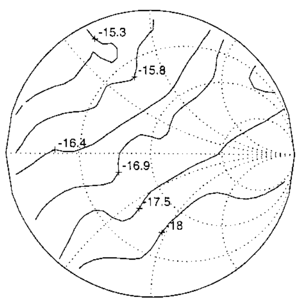


Fig. 4. Optimization of the RF source impedance. $P_L^{\rm IF}$ (3.900) in dBm versus $\Gamma_s^{\rm RF}$ (6.125).

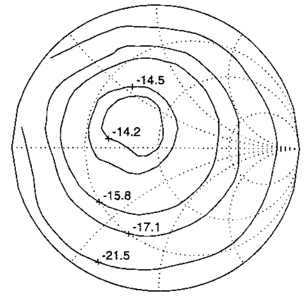
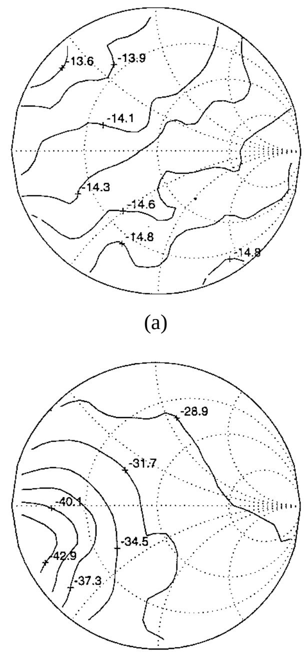


Fig. 5. Optimization of the IF load impedance. $P_L^{\rm IF}$ (3.900) in dBm versus $\Gamma_L^{\rm IF}$ (3.900).



(a)

(b)

Fig. 6. Optimization of the LO and 2LO load impedances. (a) $P_L^{\rm IF}$ (3.900) in dBm versus $\Gamma_L^{\rm LO}$ (2.225). (b) $P_L^{\rm 2LO}$ (4.450) in dBm versus $\Gamma_L^{\rm LO}$ (2.225).

(2.225 GHz) $= 0.008\angle 65^\circ$, $\Gamma_L^{\rm IF}$ (3.900 GHz) $= 0.456\angle 146^\circ$, $\Gamma_L^{\rm 2LO}$ (4.450 GHz) $= 0.550\angle{-51}^\circ$, and $\Gamma_L^{\rm LO}$ (6.125 GHz) $= 0.302\angle 180^\circ$. Fig. 3(a) and (b) show, respectively, the absorbed powers $P_L^{\rm IF}$ and $P_L^{\rm 2LO}$ as a function of $\Gamma_s^{\rm LO}$. It was observed during measurements that the level of spurious response also depends on $\Gamma_s^{\rm LO}$. Based on the above measurements, the LO source impedance is chosen to be $\Gamma_s^{\rm LO} \cong 0.7\angle{-2}^\circ$, which is a good compromise for the conversion loss, the output power level $P_L^{\rm 2LO}$, and the spurious responses.

3) Optimization of the RF Source Impedance: Active source–pull measurements are performed at the drain of the transistor for RF $=$ 5.825, 6.125, and 6.425 GHz. Fig. 4 shows a typical $P_L^{\rm IF}$ result obtained at 6.125 GHz. An optimum source impedance is found at $\Gamma_s^{\rm RF}$ (6.125) $\cong 0.92\angle 150^\circ$. The behavior of the transistor is similar at the other two frequencies, with a rotation of all constant $P_L^{\rm IF}$ contours around the center of the Smith chart of approximately $+10^\circ$ at 5.825 GHz and -15° at 6.425 GHz. It was also observed that the reflection coefficient $\Gamma_S^{\rm RF}$ does not affect the output $P_L^{\rm 2LO}$ and higher order products of the LO and RF.

4) Optimization of the IF Load Impedance: Active load–pulling is used to synthesize the IF load over the Smith chart at 3.600, 3.900, and 4.200 GHz. The RF source impedances are kept constant at $P_L^{\rm IF}$ (5.825 GHz) $\cong 0.92\angle 135^\circ$, $\Gamma_s^{\rm RF}$ (6.125 GHz) $\cong 0.92\angle 150^\circ$, and $\Gamma_s^{\rm RF}$ (6.425 GHz) $\cong 0.92\angle 170^\circ$. Fig. 5 presents the contours of constant $P_L^{\rm IF}$ (3.900 GHz) as a function of $\Gamma_L^{\rm IF}$, with the optimum load for $P_L^{\rm IF} = -14.2$ dBm is located in the area of $0.20\angle 180^\circ$. The transistor approximately shows the same behavior at 3.600 and 3.900 GHz. The load–pull measurements with the value $\Gamma_s^{\rm RF}$ (6.125 GHz) $= 0.30\angle 180^\circ$ (nonoptimum) yield the same shape of curve as in Fig. 5, but the maximum output power $P_L^{\rm IF}$ (3.900) decreases to -15.7 dBm. From these measurements, a conclusion may be drawn that a well-chosen $\Gamma_s^{\rm RF}$ can help to reduce the conversion loss of the FET resistive mixer without affecting its general behavior.

5) Optimization of the LO and 2LO Load Impedances: Fig. 6(a) and (b) presents the absorbed powers $P_L^{\rm IF}$ (3.900

TABLE I
CONVERSION LOSS VERSUS THE VARIATION IN $P_s^{\rm LO}$ AND $P_s^{\rm RF}$

Variation in P_s^{LO}			Variation in P_s^{RF}		
P_s^{LO} (dBm)	P_s^{RF} (dBm)	Loss(dB)	P_s^{LO} (dBm)	P_s^{RF} (dBm)	Loss(dB)
13.43	-5.930	-5.662	5.340	-5.930	-8.613
11.57	-5.930	-6.013	5.340	-7.710	-8.739
10.29	-5.930	-6.390	5.340	-9.440	-8.846
8.780	-5.930	-6.721	5.340	-11.31	-8.513
7.170	-5.930	-7.746	5.340	-13.30	-8.323
5.340	-5.930	-8.855	5.340	-14.84	-8.817
3.160	-5.930	-10.75	5.340	-16.91	-8.522
1.490	-5.930	-12.83	5.340	-18.96	-8.291
-0.480	-5.930	-15.17	5.340	-21.02	-8.290
-2.360	-5.930	-17.28	5.340	-23.03	-8.391
-4.380	-5.930	-19.29	5.340	-24.85	-8.460
-6.420	-5.930	-21.84	5.340	-26.86	-8.381
-8.450	-5.930	-24.02	5.340	-28.89	-8.526
-10.46	-5.930	-26.00	5.340	-30.93	-8.513
-12.34	-5.930	-28.02	5.340	-32.97	-8.550
-14.40	-5.930	-30.00	-	-	-
-16.37	-5.930	-32.13	-	-	-
-18.45	-5.930	-34.21	-	-	-

GHz) and $P_L^{\rm 2LO}$ (4.450 GHz) as a function of $\Gamma_L^{\rm LO}$ (2.225 GHz). The load impedance $Z_L^{\rm LO}$ has less effect on the power $P_L^{\rm IF}$ (3.900 GHz). To obtain significant reduction of the power $P_{\rm out}^{\rm LO}$ at the output, one can choose $\Gamma_L^{\rm LO}$ (2.225 GHz) $= 1\angle 145°$, which also minimizes the conversion loss. However, a close examination of Fig. 6(b) suggests $\Gamma_L^{\rm LO}$ (2.225 GHz) $= 1\angle 165°$. With such $\Gamma_L^{\rm LO}$, the conversion loss increases only by 1 dB, but the $P_L^{\rm 2LO}$ (4.450 GHz) can be reduced by 10 dB, which obviously offers a better tradeoff between linearity and conversion loss. It is also found that $\Gamma_L^{\rm 2LO}$ (4.450 GHz) has virtually no effect on the general behavior of the mixer in terms of the conversion loss, power $P_{\rm out}^{\rm LO}$ at the output, or higher order LO and RF products. Thus, the 2LO signal at the output can be cut down by the use of a highly reflective load with an arbitrary phase.

6) Tuning the Termination Impedances and Varying the LO and RF Source Power Levels: A final tuning showed that the termination impedances derived were near their optimum values, since no significant improvement was observed in the tuning process. Table I gives a summary of the conversion-loss variation versus the LO and RF input powers. Notice that the conversion loss varies linearly with respect to $P_s^{\rm LO}$ and it remains quasi-constant over the variation of $P_s^{\rm RF}$, which proves the good linearity of the conversion properties of the mixer.

Discussion: By moving the gate biasing point toward the pinchoff voltage, the conversion loss can be reduced from -35 dB to approximately -10 dB, with the occurrence of some higher orders of LO and RF single-tone intermodulation products. However, source–pull/load–pull single-tone measurements performed by the test set found that the resistive mixer offers good linearity of frequency conversion and, thus, low IBIM products, providing that the termination impedances are well chosen. It was found that the LO source impedance affects the conversion loss, the output power levels of the LO frequency and its harmonics, and also spurious responses. The last two characteristics are particularly important since they can also contribute to generating unwanted IM products when the mixer operates under multitone excitation. The RF source impedance, in contrast, mostly affects the conversion loss via a classical matching problem. This observation is consistent, since the intermodulation caused by the RF signal is negligible relative to that caused by the LO signal.

B. Two-Tone Measurements

The measurements were performed with a two-tone RF signal composed of the RF1 $=$ 6.1250 GHz and RF2 $=$ 6.1255 GHz. The LO and RF source impedances were set at their optimum values and only load–pull measurements were performed at the output at the LO and IF frequencies. Fig. 7 points out some hardware modifications needed to adapt the test set to two-tone measurements. First, the second harmonic 2LO load–pull branch is replaced by a second spectrum analyzer SA#2. Connected to the DUT output via a 16-dB directional coupler, this SA#2 detects a larger signal than that detected by the SA#1. As far as the terminating impedances

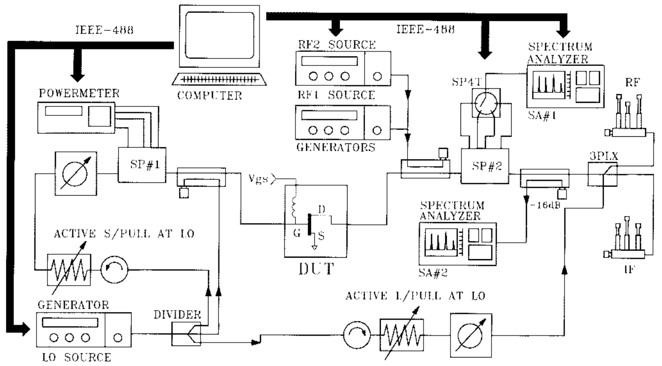


Fig. 7. Modification of the setup for the two-tone test.

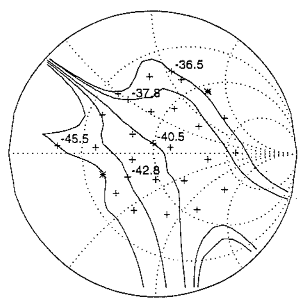


Fig. 8. 3IM/IF constant contours (in dBc) versus $\cdot +, *$: measurement points taken on the Smith chart.

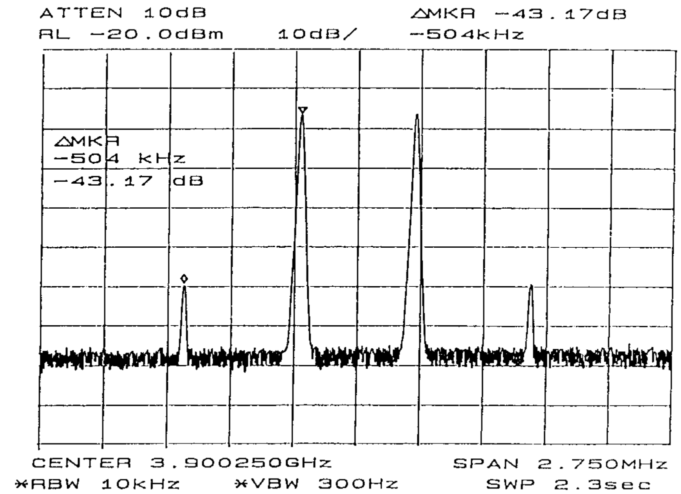


Fig. 9. Spectra of the output signal (offset +16 dB).

are concerned, the approximations and $Z_s^{\mathrm{RF1}} \cong Z_s^{\mathrm{RF2}}$ and $Z_s^{\mathrm{IF1}} \cong Z_s^{\mathrm{IF2}}$ are used since the variation in frequency is very small. Although the active RF source–pull and IF load–pull branches are still able to synthesize $Z_s^{\mathrm{RF1,2}}$ and $Z_s^{\mathrm{IF1,2}}$, a simpler technique uses three mechanical tuners to obtain the LO and RF source impedances and the IF load impedances. Hence, continuous tight-fitting of Γ_s^{LO} and $Z_s^{\mathrm{RF1,2}}$ during measurements is not needed. Due to the network losses, only the areas of circle of approximately 0.73 (SP#1) and 0.59 (SP#2) unit radius circles can be covered on the Smith charts. The LO source impedance still remains fixed at its optimum $\Gamma_s^{\mathrm{LO}} = 0.70\angle{-2^\circ}$ while the RF source impedance is fixed at $0.59\angle 140^\circ$ instead of at $0.92\angle 140^\circ$. Consequently, the measured P_L^{IF} is 0.8 dB lower than that measured with $\Gamma_s^{\mathrm{RF}} = 0.92\ \angle 140^\circ$ (see Fig. 4).

1) Active Load–Pull at the LO Output Frequency: A sweep of Z_L^{LO} over the entire Smith chart showed that Γ_L^{LO} has no effect on the IBIM product performance of the mixer. The third-order-to-carrier ratio 3IM/IF is maintained quasi-constant at -40 dBc, while the fifth-order products are submerged under the noise floor (-97 dBm). The active load–pull also allows us to explore Γ_L^{LO} outside of the Smith chart. It was observed that near a region of $\Gamma_L^{\mathrm{LO}} = 1.26\angle 5.89^\circ$, the ratio 3IM/IF drops to -50 dBc. (Physically, this load corresponds to injecting to the drain of the DUT, a LO signal having specified amplitude and phase.) However, in practice, this observation is not useful because the required output matching network would be too complex.

2) Load–Pull at the IF Output Frequency: The measurement reveals that the 3IM/IF ratio varies noticeably with respect to Γ_L^{IF}. It is observed to vary gradually from -36.5 to -45 dBc along the -135° axis (see Fig. 8). Although the optimum load for low 3IM/IF ratio does not correspond to the optimum for the low conversion loss, the two are in the same region of the Smith chart (see Fig. 5), which allows a tradeoff for overall performance. Fig. 9 shows the spectra seen at the transistor output, with $\Gamma_s^{\mathrm{LO}} = 0.70\ \angle{-1^\circ}$, $\Gamma_L^{\mathrm{LO}} = 1\angle{-162^\circ}$, $\Gamma_L^{\mathrm{IF}} = 0.25\angle 180^\circ$, and $\Gamma_s^{\mathrm{RF}} = 0.59\angle 150^\circ$. The conversion loss is around -7 dB (this would be $-7 + 0.8 = -6.2$ dB if $\Gamma_s^{\mathrm{RF}} = 0.92\angle 150^\circ$) and the 3IM/IF ratio is measured at -43 dBc.

3) Sweep the RF and LO Source Power Level: The IF signal decreases with a slope of 1 : 1 when P_s^{RF} decreases. The third-order IM products decrease more quickly with a slope of about 4 : 1 instead of the usual slope of 3 : 1. Fig. 10 shows the behavior of the mixer when P_s^{LO} is varied. It shows that the mixer behaves more linearly when the LO level is increased. It was also observed that the conversion loss remains nearly constant for $P_s^{\mathrm{LO}} \geq 0$ dBm, and increases quickly when P_s^{LO} drops below -1 dBm.

Discussion: The IF load impedance has no influence on the output LO frequency and its harmonics present at the output. It should be emphasized that a well-chosen IF load impedance is crucial, not only in the reduction of the conversion loss, but also in the reduction of the IBIM products. Although the measurements covered only a circle of 0.55 unit radius, the IF load impedance has contributed to diminishing the 3IM/IF ratio by approximately 10 dB. As for the LO and 2LO load impedances, they affect neither the conversion loss nor the IBIM performance. It is suggested that a reactive load should be used at the LO and 2LO frequencies to eliminate these unwanted output signals. Furthermore, as the output power $P_{\mathrm{out}}^{\mathrm{2LO}}$ depends on Γ_L^{LO}, the reactive load at the LO frequency should be well chosen. This remark is in agreement with previous work on FET frequency multipliers and amplifiers [15], [16]. Regarding the power level of the LO and RF generator, some observations can be made. It is known that good conversion linearity performance can be obtained by using a small RF input signal relative to the LO

TABLE II
PERFORMANCE OF THE REALIZED FET RESISTIVE MIXER MEASURED AT = +10 dBm, = −10 dBm

Conversion loss (dB)	IF/RF iso. (dB)	RF/2LO iso. (dB)	3IM/IF (dBc)	IF return loss (dB)	RF return loss (dB)	LO return loss (dB)
- 8	32	35	-51	-25	-15	-18

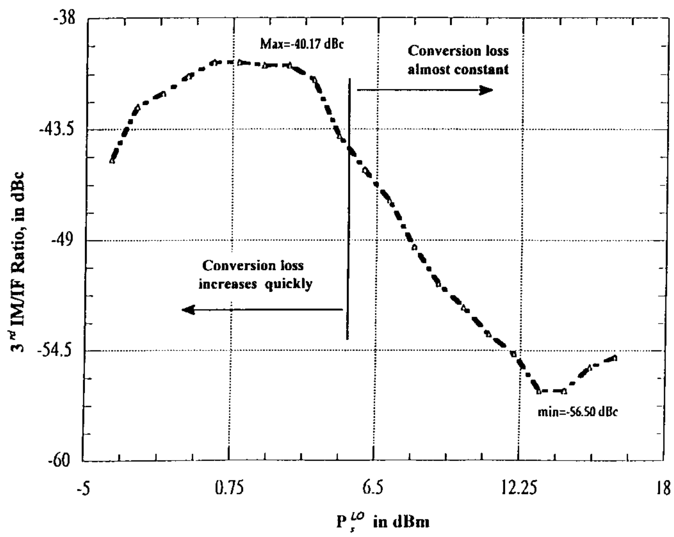


Fig. 10. 3IM/IF curve (in dBc) versus P_s^{LO}.

input level. However, using an excessively large LO signal would also deteriorate the IBIM performance. In fact, when the gate is pumped by the LO input signal, the drain-to-source resistance varies from a maximum value of $R_{\mathrm{ds}}^{\mathrm{MAX}}$ to a minimum value of $R_{\mathrm{ds}}^{\mathrm{min}}$. The constant $R_{\mathrm{ds}}^{\mathrm{MAX}}$ corresponds to the transistor pinchoff state, while the minimum $R_{\mathrm{ds}}^{\mathrm{min}}$ decreases in proportion with the increase in LO input level. When the LO positive peak voltage becomes higher than the gate built-in voltage, $R_{\mathrm{ds}}^{\mathrm{min}}$ is already at its minimum value and more nonlinearity occurs. For the NE 9001 transistor, it is found that the variation of IBIM (P_s^{LO}) level is linear when P_s^{LO} is from 4 to 12.5 dBm, and the lowest 3IM/IF ratio (−56.50 dBc) is reached at approximately $P_s^{\mathrm{LO}} = 13.2$ dBm. A good choice for the LO input level is probably 12–12.5 dBm. Generally speaking, a rule of thumb for good IBIM performance of the FET resistive mixer is to choose an LO input level such that the peak gate voltage is slightly lower than the gate built-in voltage of the transistor.

C. Design

The resistive mixer was designed and realized on a $\varepsilon_r = 10.2$ alumina substrate of 0.254-mm thickness (see Fig. 11). The overall dimensions of the circuit are 2 cm × 3 cm. Fig. 12 shows the load impedances of the output matching circuit at 2.225, 3.898, and 6.125 GHz, respectively. These load impedances are situated in the desired area on the Smith chart with respect to low IBIM products and reasonable conversion loss. Table II summarizes the performance of the realized resistive mixer. The overall performance corresponds to that observed during characterization.

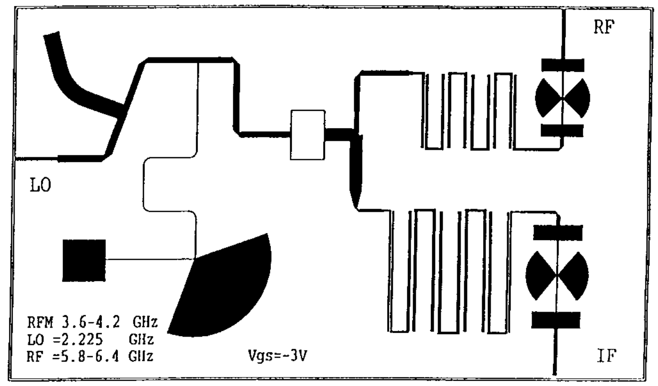


Fig. 11. Layout of the designed mixer.

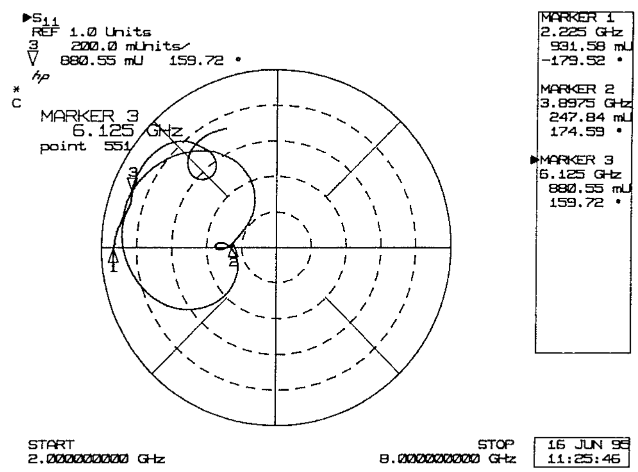


Fig. 12. Measured output matching impedances of the realized mixer.

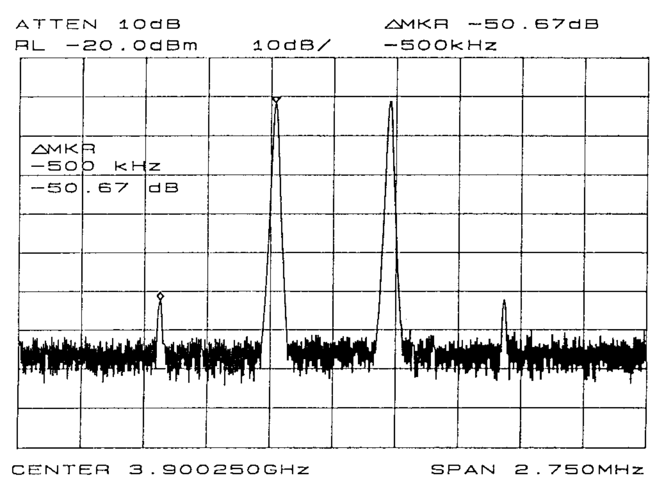


Fig. 13. Output spectra of the realized mixer (offset +16 dB).

Fig. 13 illustrates the output spectra of the mixer, measured at a higher level of $P_s^{\mathrm{RF}} = -6$ dBm and at $P_s^{\mathrm{LO}} = +10$

dBm. It should be noted that with two-tone RF excitation, the conversion loss is approximately a 10-dB loss instead of an 8-dB loss under single-tone RF excitation.

IV. Conclusion

A six-port-based measurement setup capable of synthesizing five source and load impedances simultaneously was developed to study the behavior of microwave transistors for various needs. The setup can be used to characterize transistors operated in the nonlinear mode and excited by a multifrequency signal. Active source–pull/load–pull measurements performed in such conditions allow one to study the properties of a C-band FET resistive mixer and its performance in terms of conversion loss and linearity. The full potential of the transistor can be exploited to achieve an optimum tradeoff between the loss and linearity of the frequency conversion. Furthermore, many interesting observations on the behavior of the FET under test allowed us to draw out some practical guidelines for the design of FET resistive mixers. In spite of a relatively high RF input level, it was found that excellent IBIM performance is possible with a well-chosen output IF load impedance and with sufficient LO power. Based on the results obtained, a resistive mixer was designed and built using 0.254-mm alumina substrate. The performance of the realized mixer is in good agreement with that predicted by the characterization system.

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