Large-Signal Computer-Aided Analysis and Design of Silicon Bipolar MMIC Oscillators and Self-Oscillating Mixers

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Abstract —**This paper presents a large-signal analysis and design of silicon bipolar monolithic microwave integrated circuit (MMIC) feedback oscillators and self-oscillating mixers, emphasizing the modeling of the active and passive devices and the large-signal analysis and design of nonlinear circuits using SPICE. Measured and simulated data of a *C*-band self-oscillating mixer are presented.**

I. Introduction

TIME-DOMAIN large-signal analysis programs are a very useful tool in the design of such nonlinear circuits as oscillators and mixers and in the study of the nonlinear behavior of networks in general. However, it is imperative that the active and passive device models used in these computer simulations be accurate if the analysis is to be of any value. Large-signal models for the bipolar transistor have been available for a long time and its high-frequency characteristics are thoroughly understood.

This paper presents a large-signal analysis and design of a microwave self-oscillating mixer (SOM) consisting of a silicon bipolar monolithic microwave integrated circuit (MMIC) and a dielectric resonator (DR). To illustrate the analysis methodology a television receive-only (TVRO) SOM functioning as a frequency down-converter was designed and realized [1], [2]. In a TVRO converter, the RF band from 3.7 to 4.2 GHz is down-converted to an IF band of 0.95–1.45 GHz with a high-side local oscillator (LO) signal at 5.15 GHz.

The first step in the design of a SOM is the design of the circuit as an oscillator. Fig. 1 shows the basic two-port feedback oscillator configuration. The MMIC is the active device that provides gain, and the dielectric resonator is the passive network in the feedback loop that determines the frequency of oscillation. For the circuit to oscillate, the amplitude and phase Barkhausen criteria must be satisfied.

Since the natural mode of operation of the bipolar transistor is as a current amplifier, with the appropriate form of feedback the oscillator may be said to be current controlled. Therefore the natural mode of the analysis can use current factors to satisfy the Barkhausen criteria, as an alternative to voltage factors or S parameters. The dielectric resonator couples to the transmission lines through the magnetic fields, and since there exists a close relationship between magnetic fields and current in microstrip lines, the resonator will also have a natural current transfer function.

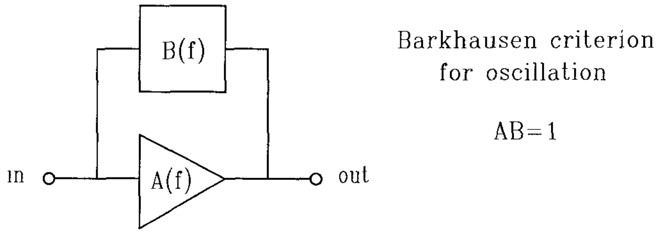

Fig. 1. Two-port feedback oscillator.

Once the circuit oscillates, if a signal is present at the input of the SOM (the base of the input transistor of the MMIC), the signal mixes with the locally generated oscillation (LO signal), generating sum and difference products. Designing the SOM adds one degree of complexity to the analysis of the oscillator because in most instances the power level of the input signal (i.e., the RF) is much smaller than that of the LO signal. Since the dynamic range of the signals involved is very large, the simulator has to "dig out" the RF signal (at about -50 dBm) from the LO (at $+10$ dBm); for intermodulation analysis the complexity increases.

This large dynamic range requires appropriate manipulation of the program tolerance limits and care to allow transients to decay to adequately low levels [3]. Furthermore, there is a high probability that the truncation interval of the time-domain analysis will not be an integral multiple of the period of oscillation, creating "leakage" when the discrete Fourier transform is performed [4]; this effect will be reviewed in detail later.

In Section II the MMIC will be described. Fabrication process details and large-signal model and performance of the MMIC will be presented. The equivalent circuit of the

Manuscript received January 2, 1988; revised August 29, 1988.
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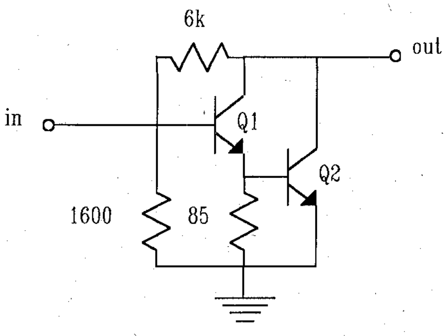
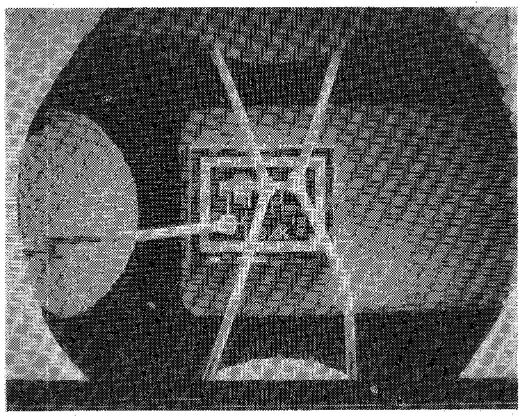
IEEE Log Number 8825372.

0018-9480/89/0300-0558$01.00



(a)



(b)

Fig. 2. (a) MMIC equivalent circuit. (b) MMIC mounted in a 70 mil stripline package.

dielectric resonator and its large-signal analysis will be discussed in Section III. Section IV covers the large-signal design of the oscillator and the details of the Fourier analysis. The simulated and measured results of the SOM are reported in Section V.

II. MMIC

Fig. 2(a) shows the equivalent circuit of the MMIC. It consists of a silicon bipolar Darlington pair and bias resistors. As a gain block the use of a Darlington pair has several advantages over a single transistor at microwave frequencies. One is that it has current gain at higher frequencies, thus significantly extending the upper limit of the frequency at oscillation. Second, a properly sized and biased Darlington pair will have a lower reflection coefficient at microwave frequencies than a single device, facilitating the matching of the device and allowing operation over a broader frequency band.

The MMIC was fabricated using an $f_T = 10$ GHz, $f_{MAX} = 20$ GHz, nitride self-aligning process featuring interdigitated 0.75-μm-wide arsenic-doped emitters, 4 μm emitter to emitter pitch, 2-μm-thick local oxide isolation, ion implantation, thin-film polysilicon resistors, and gold metallization. The small die size (0.3 mm $\times$ 0.35 mm) and the single bias supply requirement of the MMIC allow compatibility with standard microwave transistor packages. Fig. 2(b) shows a microphotograph of the chip mounted in a 70 mil stripline package.

	Q1	Q2	
Rb1	2.4	1.2	ohms
Rb2	8.7	3.1	ohms
Rb3	7.5	2.7	ohms
Rc	10	10	ohms
Re	0.7	0.3	ohms
Dbc1	377	782	area
Dbc2	225	629	area
Dbc3	130	365	area
Q	150	420	area

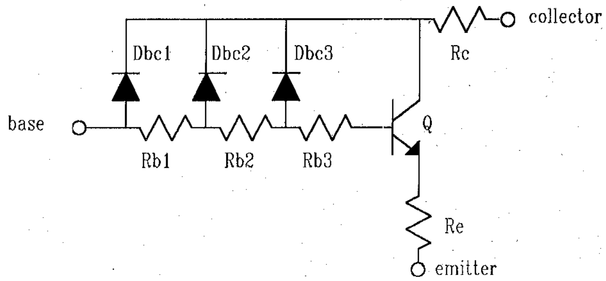


Fig. 3. Large signal transistor model.

TABLE I
SPICE MODEL PARAMETERS

Intrinsic transistor	
BF	90
IS	1.5E-18 A/μm^2
VA	20 V
IK	0.13 mA/μm^2
XTB	1.8
ISE	5.0E-15 A/μm^2
NE	2.5
TF	12 ps
PTF	40 deg
XTF	4
ITF	0.3 mA/μm^2
VTF	6V
CJE	2.4 fF/μm^2
PE	1.02 V
ME	0.6
Base-collector diode	
CJO	0.24 fF/μm^2
VJ	0.76 V
M	0.53

The large-signal model used for each transistor is shown in Fig. 3. In addition to the intrinsic transistor that simulates the action under the emitter, there is a resistor–diode ladder network to model the distributed characteristic of the extrinsic base, parasitic capacitors, and contact resistors. The intrinsic transistor uses the extended unified Gummel–Poon model in SPICE. The parameters for the model (Table I) were deduced by matching measured and simulated S parameters for single transistors under different bias conditions. Special attention was given to the high-frequency parameters, such as those that modify the forward transit time with bias. It is also important to model the parasitic effects due to resistors, metal lines, and bond wires.

The values of the resistors in the MMIC are made very high to avoid negative feedback, which would reduce the

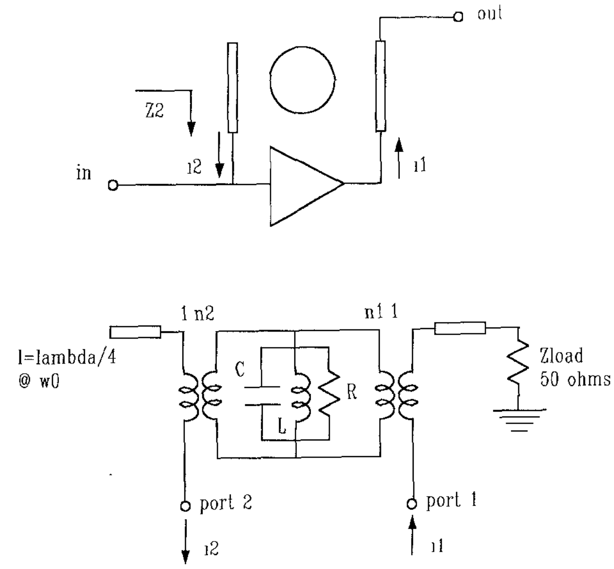


Fig. 4. SOM configuration and feedback network equivalent circuit.

nonlinearities of the transistors. At the frequency of interest for the LO (5 GHz), the current gain of the MMIC has a magnitude of 2 and a phase of 30°.

III. Dielectric Resonator Analysis

Dielectric resonators are widely used as the frequency-determining elements in feedback oscillators [5], [6]. The DR is placed between two microstrip lines (Fig. 4), one connected to the input and one to the output of the active device, coupling power from one line to the other at its resonant frequency. Complete models for transmission-type DR's coupled between lines have been presented [7].

In the model, the ideal transformers simulate the magnetic coupling of the resonator to the lines, and the *RLC* tank circuit models the resonance. The turns ratios of the transformers are a function of the distance from the lines to the resonator. The distances do not have to be equal; i.e., the DR need not be symmetrically placed between the lines. The values of the elements of the tank circuit depend on the frequency of resonance and the Q of the DR in the enclosure.

The current transfer function at resonance for the feedback network is given by (see Fig. 4)

$$\left|\frac{i_2}{i_1}\right| = \frac{n_2}{n_1}\frac{1}{1+\dfrac{n_2^2|Z_2|}{R}} \tag{1}$$

where n_1 and n_2 are the turns ratios of the transformers, Z_2 is the load impedance at port 2, and R models the unloaded loss of the resonator. For a given amplifier current gain, the ratio i_2/i_1 is designed to satisfy the Barkhausen amplitude criterion. The relationship between the current transfer function given by (1) and the S parameters of the feedback two-port shown in Fig. 4 is given by

$$\left|\frac{i_2}{i_1}\right| = \frac{2Z_0|S_{11}||S_{21}|}{Z_0|S_{21}|^2+|Z_2|[2|S_{11}|-2|S_{11}|^2-|S_{21}|^2]} \tag{2}$$

where

$$S_{11} = \frac{\beta_1}{1+\beta_1+\beta_2} \tag{3}$$

$$S_{21} = S_{12} = \frac{\sqrt{2\beta_1\beta_2}}{1+\beta_1+\beta_2} \tag{4}$$

$$S_{22} = \frac{\beta_2-(1+\beta_1)}{1+\beta_1+\beta_2} \tag{5}$$

and the coupling coefficients β_1 and β_2 are defined as

$$\beta_1 = \frac{R}{2n_1^2Z_0} \qquad \beta_2 = \frac{R}{n_2^2Z_0}.$$

Equation (2) can be used to obtain or verify the feedback current transfer ratio from measured S parameters of the transmission-mode dielectric resonator network. Conversely, if the values of n_1, n_2, and R are obtained from the analysis, (3)–(5) may be useful as a guide to experimentally establish the proper position of the resonator on the substrate.

There are some fine points involved in performing the time-domain analysis of this circuit that are worth reviewing in detail. Any second-order *RLC* network, such as the one being used to model the resonator, with $Q > 0.5$ (underdamped case), will exhibit an exponentially decaying transient response with a time constant equal to $2Q/w_0$ [8]. If, for example, a tank circuit with a Q of 350 resonating at 5 GHz is assumed, the time constant will be approximately 22 ns, and the transients will decay to adequately low levels in about 67 ns (less than 5 percent error in three times the time constant).

On the other hand, a network with this frequency selectivity will have a 3 dB bandwidth of only 14 MHz (i.e., the upper and lower 3 dB frequencies being 4.993 GHz and 5.007 GHz with time periods of 200.28 ps and 199.72 ps, respectively). Here is where the tolerance limits in the simulator program have to be carefully adjusted to avoid losing accuracy. One of the most important parameters of the time-domain analysis is the internal time step, which is the inverse of the sampling rate. The simulator may lose accuracy evaluating a signal whose period must definitely fall between 199.72 and 200.28 ps if the time step is not small enough. The particular version of SPICE that we use has a continuously adjustable internal time step to maintain accuracy. It was not accurate enough, however, to keep the signal within the necessary frequency selectivity that the filter required, and we had to force a very short time step even after modifying other tolerance limits. We found out empirically that the internal time step has to be limited to 2 ps for SPICE to correctly simulate a 5 GHz second-order *RLC* network with a Q of 350.

From a pragmatic point of view, the designer now faces a dilemma: to accurately simulate this filter with a loaded Q of 350 around 5 GHz, the internal time step has to be limited to about 2 ps. But if the transients alone require 67 ns to decay, 33 500 samples are required just to calculate the transients. This dilemma is illustrated in Fig. 5, which

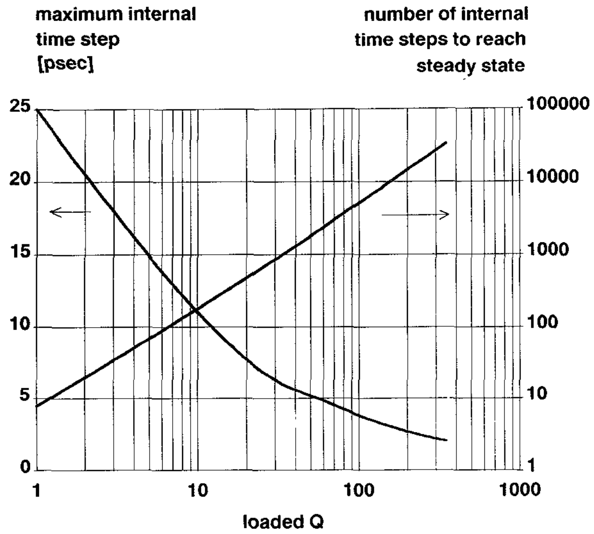


Fig. 5. Maximum internal time step and number of internal time steps to reach steady state versus loaded Q for a second-order RLC network resonating at 5 GHz.

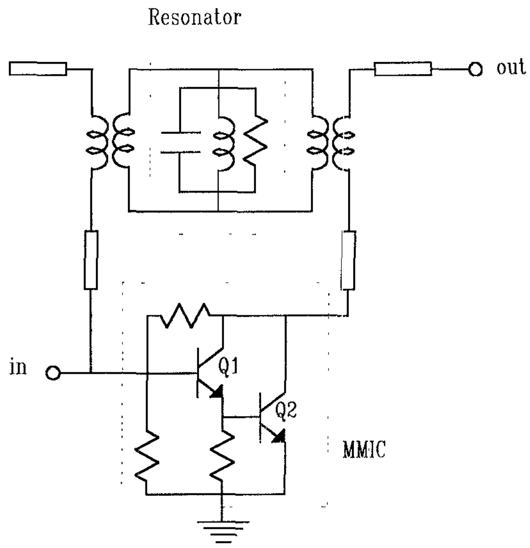


Fig. 6. SOM equivalent circuit.

shows a plot of maximum internal time step and number of internal steps to reach (within 5 percent) steady state as a function of Q for a second-order RLC network resonating at 5 GHz.

Although the circuit under investigation is relatively simple, when the complete SOM is simulated, 60 nodes are involved and the time for each simulation becomes unacceptably long. To reduce the resonator's transient time and allow a larger internal time step, we decided to artificially lower the loaded Q to 15 and found out empirically that the internal time step had to be limited to about 9 ps (Fig. 5) to correctly simulate the network. Under these conditions, using a 20 MHz 80386-based computer the simulation time for the complete SOM was approximately 25 minutes for a 50-ns-long transient analysis. Since our main objective was to model the conversion gain of the SOM, and we were not particularly interested in simulating frequency stability or transient response, the lowering of the Q did not have a significant impact. The simulated conversion gain of the SOM changed less than 0.15 dB when varying the loaded Q from 15 to 75 (see Fig. 7). It should be emphasized that after the values for the RLC elements and the turns ratios of the transformers have been found from a small-signal analysis, a time-domain analysis of the feedback network should be performed to guarantee that the tolerance limits and the internal time step are correctly set, and that the transients are decaying to an acceptable level.

IV. Oscillator Analysis

Once both the active device and the frequency-determining element have been analyzed separately, an oscillator satisfying the Barkhausen criteria can be designed. The complete equivalent circuit of the SOM is shown in Fig. 6. The DR's transfer function (1) and the length of the transmission lines are adjusted to satisfy the amplitude and phase criteria, respectively.

After these adjustments are made, a second large-signal simulation of the open-loop circuit should be performed to guarantee that the tolerance limits are properly set for the complete SOM. SPICE has some difficulty with the transmission line model in the time-domain analysis at high frequencies when long analyses are performed. Therefore, after obtaining the required length of line from the small-signal simulation, an LC ladder was used to model the lossless lines.

The next step in the design is to close the loop and apply a short transient stimulus to the device to trigger the oscillation—much the same role that noise plays in actual circuits. Fig. 7 shows output voltage waveforms of the oscillator for loaded Q values of 15, 30, and 75. The buildup times for oscillation are about 10, 20, and 50 ns, respectively. The buildup time for oscillation is directly proportional to the loaded Q of the circuit and for this particular example is approximately ten times the time constant of the RLC resonator. The LO power to the output 50 Ω load is +7 dBm.

The analysis may also be used to determine the limiting mechanism in the oscillator. Transistor current waveforms (Fig. 8) indicate that transistor Q1 is the device turning off, while Q2 has a linear response. From these data and from small-signal current gain information, one can deduce that Q1 is the nonlinear device limiting the amplitude of oscillation and Q2 acts as a current amplifier.

The output of the time-domain analysis was then Fourier transformed. Note that when obtaining the discrete Fourier transform of the time-domain signal, if the truncation interval (e.g., final time value of the analysis) of the sampled signal is not an integral multiple of its period, the resulting discrete frequency spectrum will exhibit side lobes [4]. These side lobes are responsible for the additional frequency components, termed leakage. The problem appears when performing the convolution in the frequency domain of a sinc(f) function with zeros at multiples of $1/T_{\text{trunc}}$ (which is the Fourier transform of a rectangular

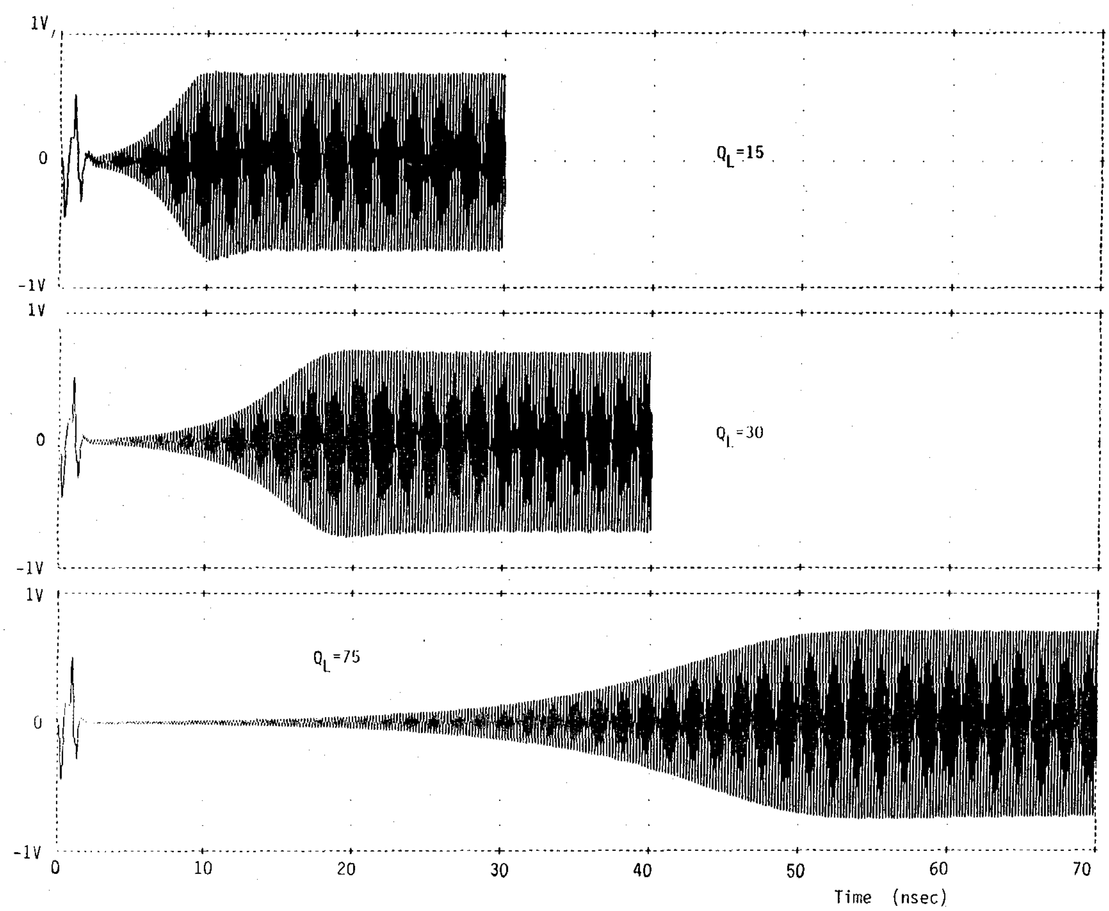


Fig. 7. Oscillator output voltage waveforms for loaded Q values of 15, 30, and 75.

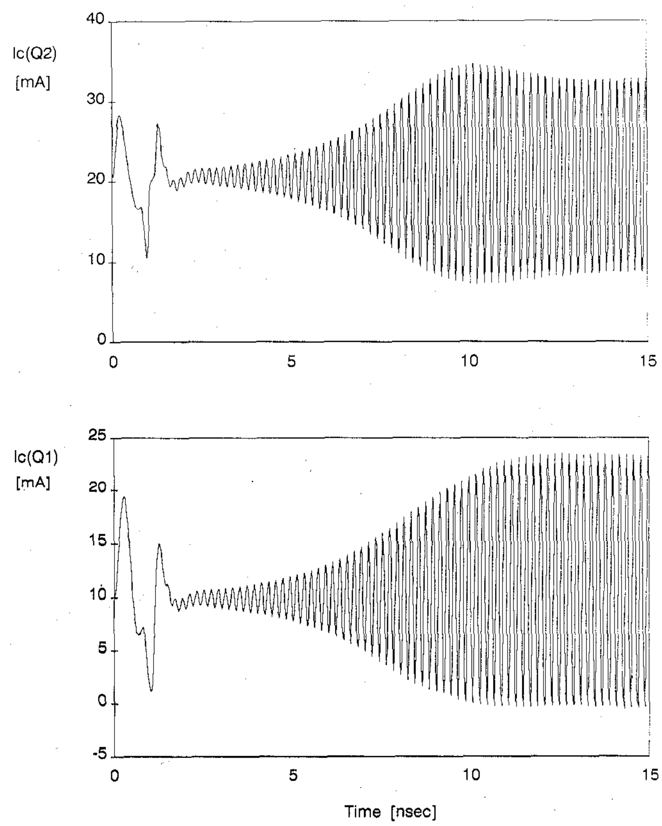


Fig. 8. Collector current waveforms for transistors Q1 and Q2.

truncation function) with the Fourier transform of the signal of interest having a period $1/T_0$, when T_{trunc} is not an integral multiple of T_0. To reduce leakage it is necessary to employ a time-domain truncation function which has side lobe characteristics that are of smaller magnitude than those of the $\text{sinc}(f)$ function. The smaller the side lobes, the less leakage will affect the results of the transform. One particularly good truncation function is the Hanning function [4] (also known as the raised cosine function in communication systems), given by

$$h(t) = 1/2 - 1/2 \cos \frac{2\pi t}{T_{\text{trunc}}}, \qquad 0 < t < T_{\text{trunc}}.$$

Fig. 9 shows the discrete Fourier transform of the oscillator's output signal, using both the regular rectangular truncation function and the modified (Hanning) function. The leakage reduction is of the order of 50 dB. There is, however, one disadvantage to using the Hanning function. To get the same frequency step in the frequency response the truncation interval of the Hanning function has to be twice as long as when using the rectangular function.

For oscillator design, where only the amplitude and frequency of oscillation are of interest, the leakage may be of no consequence; however, for SOM's the RF and IF signals could well be buried in the leakage and it becomes of the utmost importance to keep its level as low as possible.

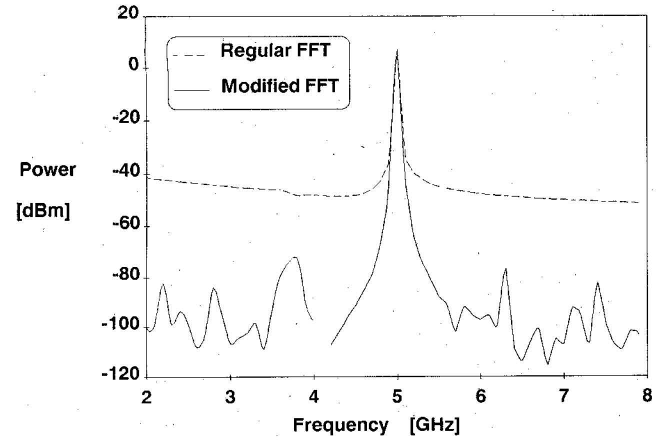


Fig. 9. Oscillator power spectrum.

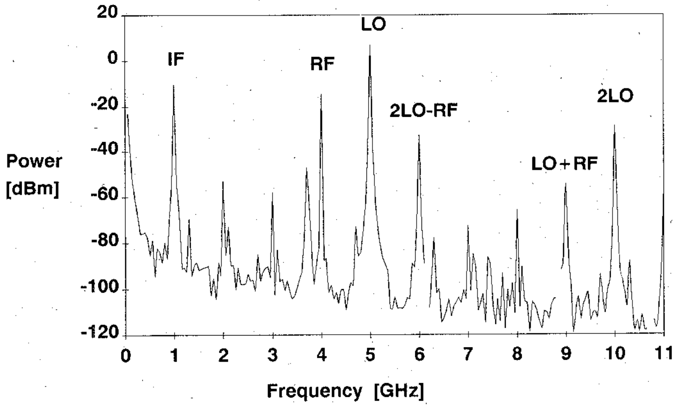


Fig. 10. SOM output power spectrum.

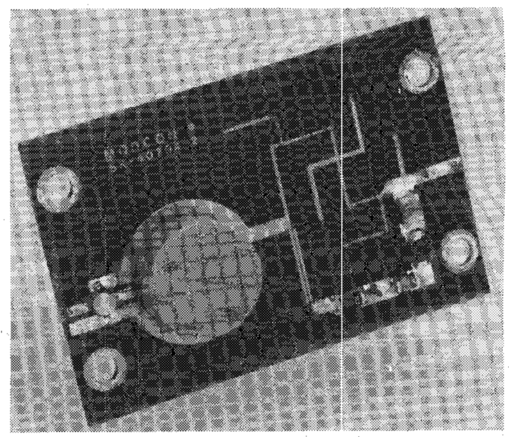


Fig. 11. TVRO SOM board including MMIC, DR, bias, and output filter.

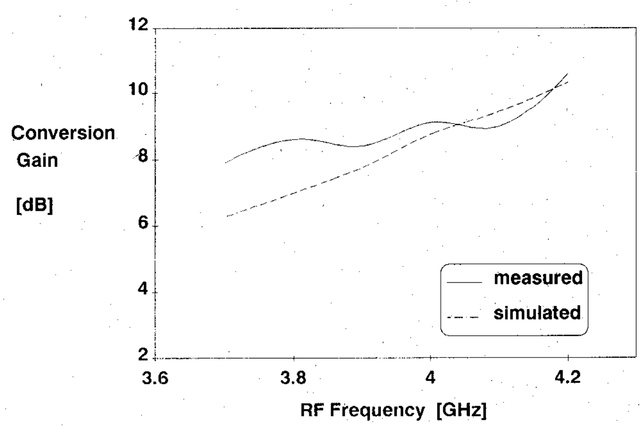


Fig. 12. Measured and simulated performance of TVRO SOM.

V. Self-Oscillating Mixer

To simulate the performance of the SOM, a 4 GHz signal was applied at the input of the SOM, and the output signal was analyzed. Fig. 10 shows the spectrum at the output port. Since the SOM is not a balanced network, there is no isolation between the ports. For the TVRO band, the simulation predicts 9 ± 1.5 dB of conversion gain. We also obtained the frequency spectrum of the currents in Q1 and Q2, and concluded from the small-signal current gain information that Q1 is the nonlinear device generating the mixing products and Q2 acts as an output amplifier. An advantage of the Darlington pair over a single transistor in a SOM is that the bias of each device is set independently, which provides independent control over the amplitude of oscillation and the conversion gain. This can offer advantages in noise figure and/or distortion performance.

An experimental prototype was fabricated using a 31-mil-thick epoxy-glass (FR4) board (dielectric constant = 4.8). The MMIC was packaged in a 70 mil microstrip ceramic package and mounted on the board as shown in Fig. 11. Plated through-holes directly under the two ground leads on the package were used to ensure proper grounding.

Using a DR with a resonant frequency of 5.15 GHz, a dielectric constant of 37, and an unloaded Q of 7000, a TVRO down-converter was realized. With the MMIC biased at 35 mA and 8 V (from a 15 V power supply through a 200 Ω dropping resistor), the down-converter exhibits 9 ± 1 dB conversion gain. Fig. 12 shows the good agreement between the measured and simulated conversion gain. Additionally, the simulation predicted the output compression point of the SOM to be +7 dBm, which was in agreement with the measured data with an error of less than 0.5 dB.

VI. Summary and Conclusions

The large-signal analysis and design of a silicon bipolar MMIC self-oscillating mixer using a time-domain simulator program have been presented. Detailed modeling of the active and passive devices is necessary for the analysis to be of any value. Problems simulating high-Q networks and obtaining the Fourier transform of a class of signals have been reported and some solutions given. Time-domain simulator programs are very useful tools in the design of nonlinear IC's, but their operation, specifications, and limitations must be understood to effectively evaluate the results.

Acknowledgment

The authors thank Dr. C. P. Snapp for his encouragement, guidance, and many useful suggestions during the course of this work.

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Dr. Khanna has authored or coauthored more than 30 publications and four books. He holds five U.S. and two French patents on dielectric resonator circuits.