COMPUTER SIMULATION MODELS PERFORMANCE OF RF CONVERTERS

Self-oscillating mixers can be accurately designed using time-domain analysis and SPICE.

LARGE-signal analysis is crucial for accurately modeling self-oscillating mixers or frequency converters. However, time-domain simulation programs such as SPICE require detailed models and careful control of tolerance limits. Fortunately, large-signal transistor models are readily available, and valuable results can be obtained if the simulation program is understood.

When designing a self-oscillating microwave mixer, the first step is to consider the circuit as an oscillator. In a basic two-port feedback oscillator, a bipolar MMIC amplifier can provide gain, and a dielectric resonator in the feedback loop can determine the frequency of oscillation. The amplifier and the feedback element have frequency-dependent transfer functions A(f) and B(f), respectively. For the circuit to oscillate, the Barkhausen criterion (AB = 1∠0°) must be satisfied.

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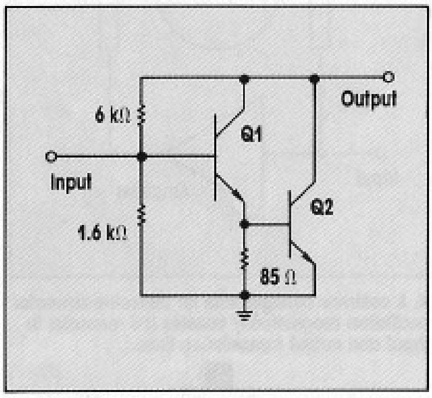

1. The MMIC amplifier used in a self-operating mixer operating with a 5.15-GHz LO utilizes two bipolar transistors connected as a Darlington pair.

Because the basic mode of operation for a bipolar transistor is that of a current amplifier, the device is readily analyzed in terms of current factors. Additionally, the dielectric resonator is magnetically coupled to transmission lines, and there is a close relationship between magnetic fields and currents in microstrip lines. Therefore, the oscillator is easily described in terms of current transfer functions.

When an RF signal is applied to the input of the self-oscillating mixer (the base terminal of the input transistor), the signal mixes with the local oscillator (LO) and produces sum and difference products. In most instances, the RF signal's power level is much less than that of the LO signal, and the dynamic range of the signals involved is 60 dB or more. For intermodulation analysis, dynamic range is even greater.

The equivalent circuit of a typical MMIC amplifier used in a self-oscil-

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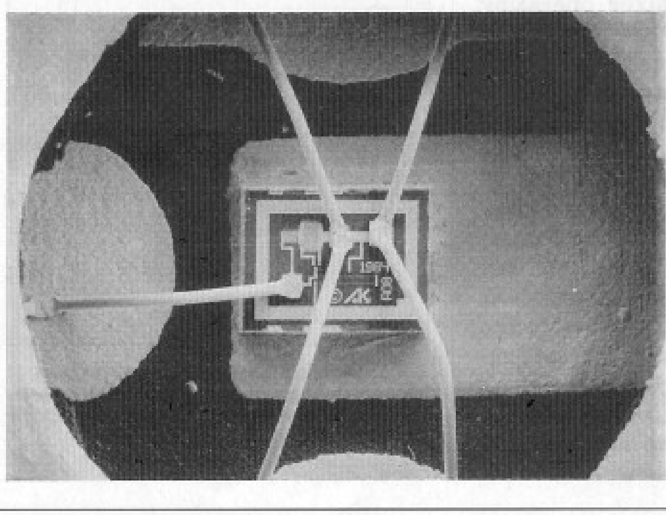
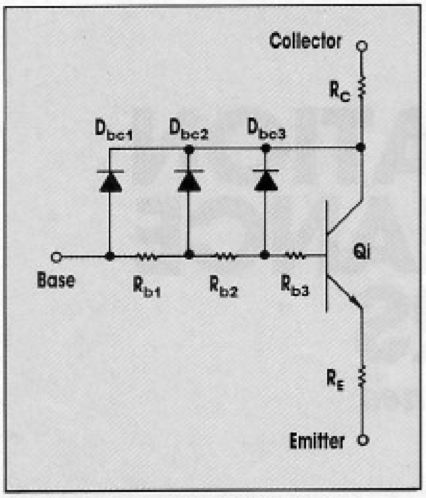


2. Because the MMIC is small and can operate from a single bias voltage, it can be mounted in a standard 70-mil stripline package.



3. Transistors are modeled with an intrinsic transistor (Qi) to simulate action under the emitter. Resistors and diodes simulate the extrinsic base and contact resistances.

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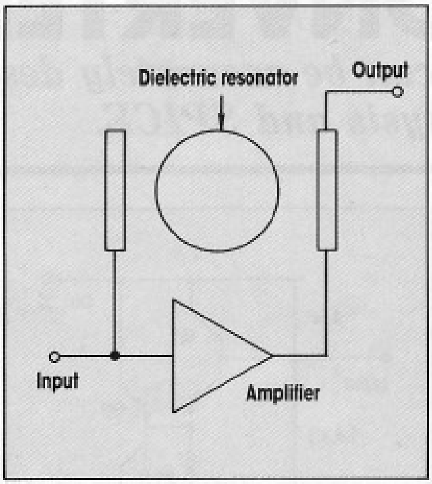
lating mixer consists of a silicon bipolar Darlington pair with bias resistors (Fig. 1). As a gain block, the Darlington pair has several advantages over a single device at microwave frequencies. It has current gain at higher frequencies, which results in an extension of the maximum frequency of oscillation. If properly sized and biased, Darlington pairs have a lower reflection coefficient than a single device, easing impedance matching and allowing operation over a broader range of frequencies.

The MMIC amplifier employed in a self-oscillating mixer operating with a 5.15-GHz LO frequency was fabricated using a self-aligning nitride process. The process produces bipolar devices with 10-GHz transition frequency (f_T) and 20-GHz maximum frequency of oscillation (f_{MAX}). Ion implantation forms interdigitated transistors with 0.75-μm emitter width, 4-μm emitter-to-emitter pitch, and 2-μm local oxide isolation.

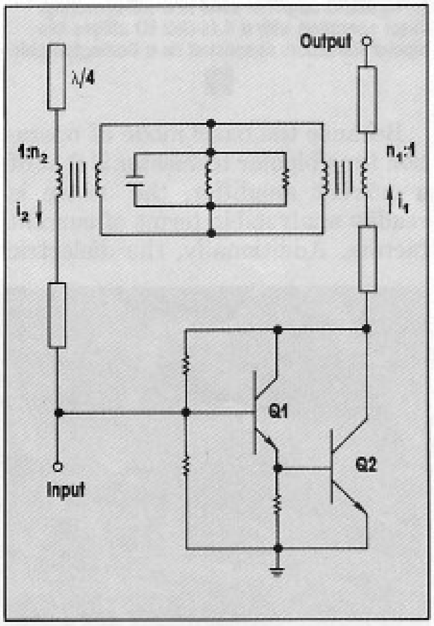
Thin-film polysilicon resistors and gold metallization form the bias circuit and interconnections. The small die size (0.30 × 0.35 mm) and the need for only one bias supply render the MMIC compatible with most mi-

Table 1: Transistor resistances (Ω)

Transistor	R_{b1}	R_{b2}	R_{b3}	R_C	R_E
Q1	2.4	8.7	7.5	10.0	0.7
Q2	1.2	3.1	2.7	10.0	0.3



4. A common configuration for dielectric-resonator oscillators magnetically couples the resonator to input and output transmission lines.



5. The complete oscillator circuit uses two ideal transformers to model magnetic coupling between the resonator and microstrip lines. The resonator is simulated by an RLC tank circuit.

crowave transistor packages, including the 70-mil stripline package that was used (Fig. 2).

The large-signal model for each transistor includes an intrinsic transistor that simulates action under the emitter (Fig. 3). Additionally, a resistor-diode ladder network models the distributed characteristics of the extrinsic base (Tables 1-2). Contact resistances are also included in the model. The intrinsic transistor uses the extended, unified Gummel-Poon model in SPICE.

Parameters for the intrinsic transistor model were deduced by matching measured and simulated S-parameters for single transistors under various bias conditions (Table 3). Special attention was given to high-frequency parameters such as those that modify the forward transit time with DC bias. Care was also given to modeling the parasitic effects of resistors, metal lines, and bond wires.

To avoid negative feedback a maximize the MMIC's nonlinear characteristics, resistances were made relatively high. At the 5.15-GHz LO frequency, the MMIC exhibits a current gain of 2 with 30-deg. phase shift.

FEEDBACK NETWORK

When used as the frequency-determining element in a feedback oscillator, a dielectric resonator is often placed between two microstrip lines[1-2] (Fig. 4). These lines are connected to the input and output of the active device, and the resonator couples power between the lines. The model used to analyze the resonator utilizes ideal transformers to simulate magnetic coupling between the resonator and the lines.[3] A resistor-inductor-capacitor (RLC) circuit models resonant behavior (Fig 5).

The turns ratio of transformers used in the model is determined by the distance between the resonator and the microstrip line. The resonator may be placed closer to one of the lines, resulting in differe turns ratios for the two transformers. Elements in the tank circuit de-

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pend on the resonant frequency and Q of the resonator. At resonance, the magnitude of the feedback network's transfer function is given by:

$$|i_2/i_1| = (n_2/n_1) \times (1 + n_2^2 |Z_2|/R)^{-1} \quad (1)$$

where:
i_1 = output current,
i_2 = input current,
n_1 = output turns ratio,
n_2 = input turns ratio,
Z_2 = impedance at port 2, and
R = RLC tank resistance.

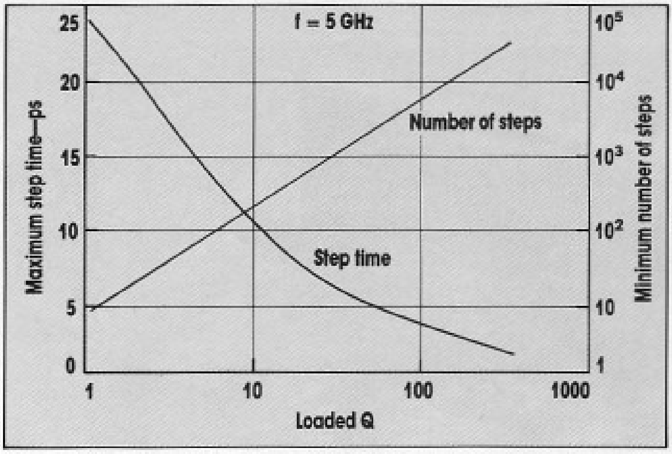
To satisfy the Barkhausen criterion, the magnitude of i_1/i_2 must equal the active device's current gain. Assuming port 1 of the feedback network is terminated with the system's characteristic impedance, the transfer function given by Eq. 1 is related to the network's S-parameters:

$$|i_2/i_1| = 2 Z_0 |S_{11}||S_{21}| \times [Z_0 |S_{21}|^2 + |Z_2|(2|S_{11}| - 2|S_{11}|^2 - |S_{21}|^2)]^{-1} \quad (2)$$

where:
Z_0 = system characteristic impedance.

The S-parameters in Eq. 2 can be related to the coupling coefficients, β_1 and β_2:

$$S_{11} = \beta_1 (1 + \beta_1 + \beta_2)^{-1} \quad (3)$$



6. As the resonator's loaded Q increases, a smaller time step is required to maintain accuracy. The circuit's time constant also increases, requiring a large number of calculations or steps.

$$S_{12} = S_{21} = (2 \beta_1 \beta_2)^{1/2} \times (1 + \beta_1 + \beta_2)^{-1} \quad (4)$$

$$S_{22} = [\beta_2 - (1 + \beta_1)] \times (1 + \beta_1 + \beta_2)^{-1} \quad (5)$$

The coupling coefficients are determined by resonator losses and turns ratios:

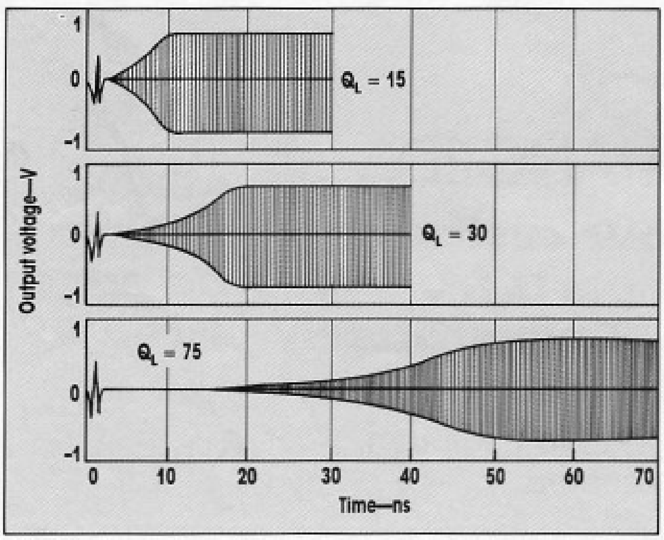
$$\beta_1 = R/2 n_1^2 Z_0 \quad (6)$$

$$\beta_2 = R/n_2^2 Z_0 \quad (7)$$

Typically, the feedback network's S-parameters are measured using a vector network analyzer, and Eq. 2 is used to determine the network's transfer function in terms of currents. Turns ratios and the resonator loss factor can also be derived. These factors may be used as guides for determining the resonator's position.

Table 2: Junction areas (μm^2)

Transistor	Qi	D_{bc1}	D_{bc2}	D_{bc3}
Q1	150	377	225	130
Q2	420	782	629	365



7. Varying the loaded Q from 15 to 75 has little effect on the oscillation's amplitude. However, the time constant increases in proportion to the loaded Q.

RESONATOR ANALYSIS

A second-order RLC circuit with a loaded Q greater than 0.5 (underdamped) will exhibit an exponentially decaying transient response.[4] The response has a time constant equal to $2Q/\omega_0$, where ω_0 is the resonant frequency in rad/s. For example, a resonator with Q = 350 at 5.0 GHz has a time constant of 22 ns. The transient response falls to adequately low levels (less than 5-percent error) in about three time constants or 67 ns. The 3-dB bandwidth of the resonator, however, is only 14 MHz. The 3-dB points are at 4.993 and 5.007 GHz, which correspond to time periods of only 200.28 and 199.72 ps, respectively.

At this point, the tolerance limits in the simulator program must be carefully adjusted to ensure accuracy.[5] One of the most important parameters of the time-domain analysis is the internal time step, which is the inverse of the sampling rate. When evaluating a signal with a period that is between 199.72 and 200.28 ps, the simulation program can lose accuracy if the time step is not small enough.

Although the version of SPICE that was used to evaluate dielectric resonators in self-oscillating mixers has a continuously variable time step to maintain accuracy, the program was not accurate enough to keep the signal within the frequency limits required by a 5-GHz resonator with Q = 350. It was necessary to use a very short time step in the analysis, regardless of any adjust-

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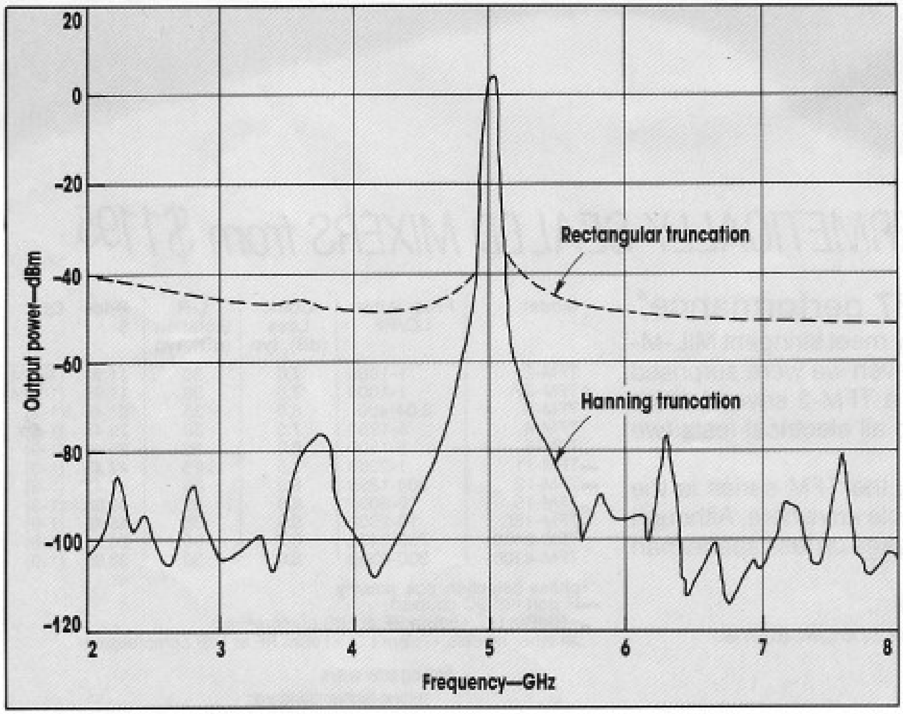
ments that were made to other tolerance limits. The internal time step necessary for sufficient accuracy was determined empirically. As noted, for a 5-GHz, second-order RLC resonator with Q = 350, the maximum internal time step is about 2 ps.

Such a resonator, however, requires about 67 ns for the transient signal to decay. Using 2-ps time steps, 33,500 samples are required to calculate the transient response alone. A plot of the maximum internal time step and the corresponding number of steps needed to reach the steady state indicates that high-Q circuits require very large numbers of calculations (Fig. 6).

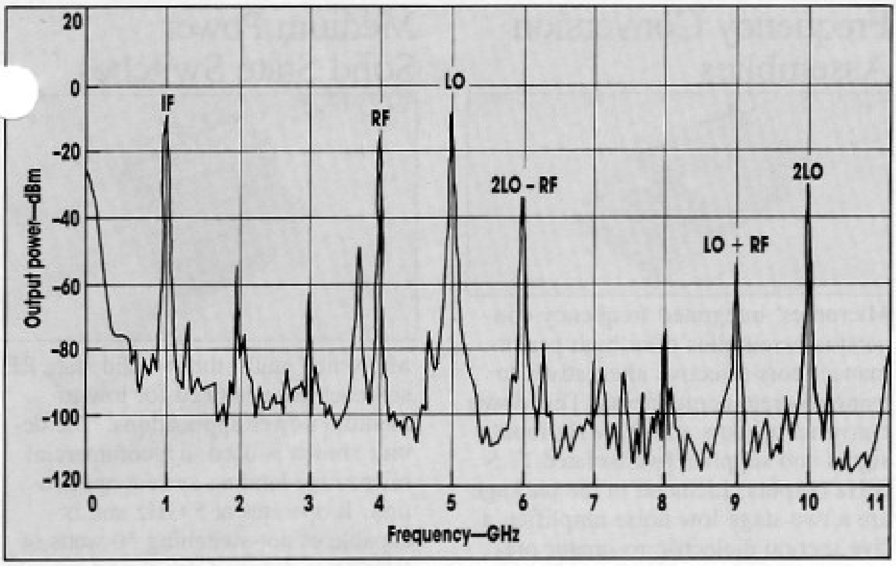
Although the self-oscillating mixer is a relatively simple circuit, it requires about 60 nodes. The time required to analyze the circuit using a resonator with Q = 350 becomes unacceptably long. To reduce the resonator's transient time and allow a larger internal time step, the resonator's Q was artificially reduced to 15. The maximum time step to achieve sufficient accuracy was thus increased to 9 ps. Analyzi the modified circuit over a 50-ns time span required 25 min. using a computer with a 20-MHz clock.



8. Using the Hanning truncation function decreases leakage terms by about 50 dB. With a rectangular truncation, signals could be buried in leakage terms.

Table 3: SPICE model parameters

Intrinsic transistor (Qi)	
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	6.0 V
CJE	2.4 fF/μm^2
PE	1.02 V
ME	0.6
Base-collector diodes	
CJO	0.24 fF/μm^2
VJ	0.76 V
M	0.53



9. With a 5.0-GHz LO frequency and a 4.0-GHz RF signal applied, a Fourier transform of the simulated mixer's output reveals IF, image, and intermodulation products.

The main objective of analyzing the self-oscillating mixer was to determine its conversion gain. Lowering the resonator's Q to 15 did not significantly affect this parameter. Simulated conversion gain typically changed by less than 0.15 dB when the resonator's loaded Q was varied from 15 to 75. Before analyzing the entire mixer, however, a time-domain analysis should be performed on the resonator to guarantee that tolerance limits and the internal time step used in the analysis provide sufficient accuracy. Additionally, transients should decay to acceptable levels.

OSCILLATOR ANALYSIS

After the active device and feedback network have been analyzed, an oscillator can be designed. In the complete equivalent circuit for the self-oscillating mixer (Fig. 5), the resonator's transfer function and the lengths of the transmission lines are adjusted to satisfy amplitude and phase requirements of the Barkhausen criterion. After these adjustments are made, a second large-signal simulation of the open-loop circuit should be performed to make sure that the mixer can be analyzed accurately using previously determined tolerance limits.

The transmission-line model in SPICE exhibits some problems when used in a long time-domain analysis at high frequencies. Therefore, after obtaining the required lengths of line from a small-signal

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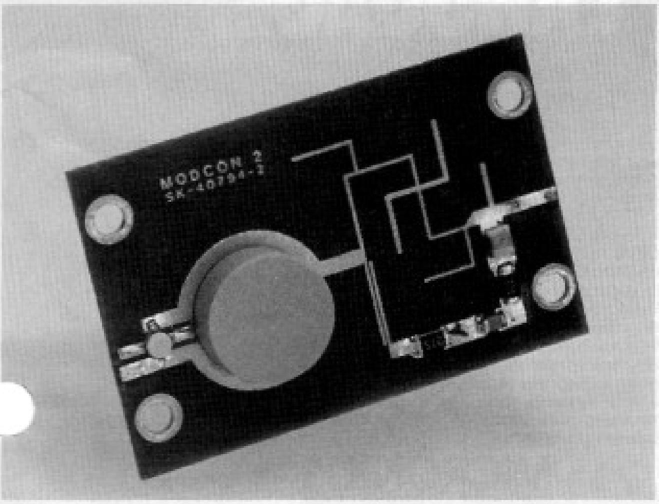

10. A prototype mixer was based on the design obtained from simulations. Constructed on a glass-epoxy board, the mixer uses a 5.15-GHz dielectric resonator with an unloaded Q of about 7000.

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simulation, the lines were substituted with lossless inductor-capacitor (LC) ladder networks. These networks include many of the 60 nodes used in the analysis.

After the oscillator loop was closed, a short transient stimulus was applied to trigger the oscillation. The stimulus performs the same role as random noise in the actual circuit. Using Qs of 15, 30, and 75, the buildup times for the oscillation were approximately 10, 20, and 50 ns, respectively (Fig. 7). Buildup time is directly proportional to the circuit's loaded Q and, for this example, is about ten times the time constant of the RLC resonator. Power delivered to the 50-Ω load is approximately +7 dBm for all three values of Q.

Transient analysis can also be used to determine the mechanism that limits the oscillation's amplitude. An analysis of transistor current waveforms indicated that the input transistor, Q1, begins to turn off as amplitude increases, while Q2 maintains a linear response. From this data and the devices' small-signal current gains, it can be deduced that Q1 is the nonlinear device that limits amplitude, and Q2 acts as a linear current amplifier.

Before analyzing the self-oscillating mixer's conversion gain, a discrete Fourier transform is performed on the oscillation to determine the best truncation interval, T_{TR} (i.e., the time span of the analysis) and to implement a suitable truncation function. If the truncation interval is not an integer multiple of the signal, the resulting frequency spectrum exhibits sidelobes.[6] These sidelobes represent leakage or undesired frequency components not generated by the circuit.

Leakage is equivalent to performing a convolution in the frequency domain between the signal and a Sinc(f) function that has zeros at multiples of $1/T_{TR}$. The Sinc(f) function is the Fourier transform of a rectangular pulse in the time domain.

To reduce leakage, a truncation function can be used. The truncation function and the signal are multiplied in the time domain before performing the Fourier transformation. If the truncation function has smaller frequency-domain sidelobes than those of a rectangular pulse, less leakage is generated. A particularly good truncation function is the Hanning function, h(t), which is also known as the raised cosine function:

$$h(t) = 0.5 + 0.5 \times \cos(2\pi t/T_{TR})$$

where:
t = time.

Although the Hanning function reduces leakage by about 50 dB (Fig. 8), the truncation interval must be twice as long as that used with the rectangular pulse to achieve the same frequency resolution. When designing an oscillator, leakage terms may have no consequence. However, because of the large dynamic range involved, RF and IF signals in a self-oscillating mixer could get buried in leakage terms. Thus, leakage terms must be minimized in an analysis of such circuits.

MIXER PERFORMANCE

With a 4.0-GHz signal applied to the input of the self-oscillating mixer, the output's frequency spectrum was simulated (Fig. 9). Because the mixer is not a balanced network, there is no isolation between ports. For the TVRO band (3.7 to 4.2 GHz), the analysis predicts approximately 9 ± 1.5-dB conversion gain. A Fourier analysis of the currents in transistors Q1 and Q2 indicated that Q1 mixes the RF and LO signals. This could be ex-

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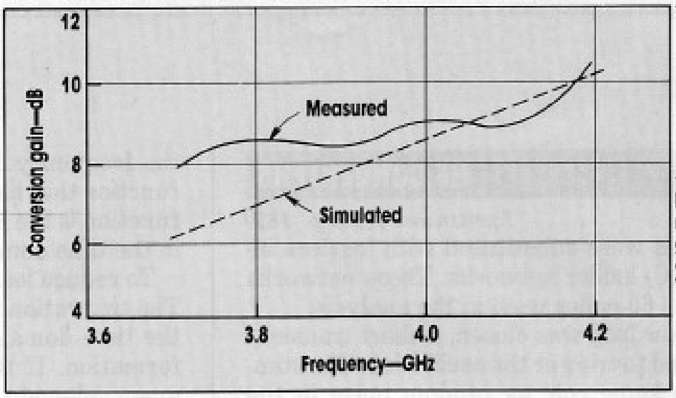
pected because Q1 operates nonlinearly.

An important advantage to using the Darlington pair is that the bias of each device can be set independently, providing separate control of oscillation amplitude and conversion gain. These controls are useful for adjusting the mixer's noise and distortion characteristics.

A prototype self-oscillating mixer was fabricated on a 31-mil-thick epoxy-glass board ($\epsilon_r = 4.8$). The MMIC was placed in a 70-mil microstrip package and was mounted to the board (Fig. 10). Plated through-holes were placed directly under the MMIC to ensure effective grounding.[7-8]

Using a 5.15-GHz dielectric resonator with $\epsilon_r = 37$ and an unloaded Q of about 7000, a TVRO downconverter was realized. With the MMIC biased with 8.0 V at 35 mA, t[he] downconverter provides 9 ± 1-dB conversion gain (Fig. 11). The output compression point was also measured and is within 0.5 dB of the +7-dBm value predicted by the analysis. Although detailed modeling and careful control of the simulation program were necessary to obtain useful results, time-domain analysis can be a powerful tool in the design of many nonlinear microwave circuits. ••



11. The difference between simulated and measured conversion gain is less than 2 dB across the 3.7-to-4.2-GHz TVRO band.

Acknowledgment

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