A Balanced Self-Oscillating Mixer

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***Abstract*—A balanced self-oscillating mixer is proposed. It consists of a pair of AlGaAs/GaAs 10×45 μm pHEMT's, oscillating at 7.53 GHz and uses the extended resonance effect. The circuit exhibits a conversion gain of 3.6 dB and reduces the second-order intermodulation products by 18.3 dB. The balanced nature of the oscillators also provides good LO to RF isolation of 40.5 dB when used as a upconverter. This approach relaxes the filtering requirements for generating single-sideband AM.**

***Index Terms*—Balanced mixers, HEMT mixers, self-oscillating mixers.**

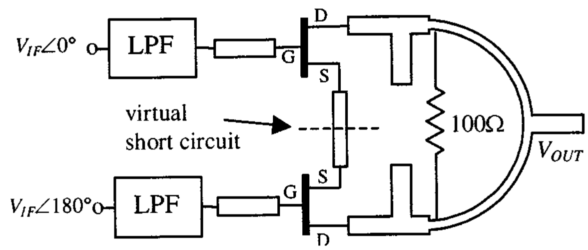

Fig. 1. Balanced SOM microstrip layout.

I. Introduction

Emerging broadband communication systems require compact, low cost front-end transceivers. This may be achieved by integrating circuit functionality within active devices, leading to lower power consumption and improved reliability. Of particular interest are the local oscillator and mixer functions, which have been combined using a self-oscillating mixer (SOM) configuration [1]–[3]. A disadvantage of the self-oscillating mixer when used as an upconverter is that it does not provide good LO to RF isolation. Although the RF and LO may be separated in frequency sufficiently for filtering to be employed, up-converting SOM's possess high conversion gain if the RF is close to the LO signal [4].

In order to provide good LO-RF isolation and reduce intermodulation, a balanced approach is proposed. Previous work on balancing a SOM used a dielectric resonator [5], which is unsuitable for MMIC fabrication. Here, a novel topology for a balanced self-oscillating mixer based on the extended resonance technique [6] is investigated for the first time. This approach allows for MMIC integration and also has reduced intermodulation compared to conventional SOM's. Stabilization of the circuit could be achieved by means of subharmonic injection locking [4].

II. Self-Oscillating Mixer Design

A pair of Agilent Technologies 10×45 μm pHEMT's were used to provide oscillation. In order to achieve LO rejection at the output, the device oscillations must be superimposed out of phase. In [6], the gates of a pair of oscillators were connected by a length of microstrip line such that a virtual short circuit is set up at the mid-point of the transmission line, generating an oscillation which is 180° out of phase. In contrast, the source terminals are connected together in this application (Fig. 1), thus allowing the gate terminals to be used as input terminals. This is similar to a "transconductance mixer" topology. The outputs of the balanced oscillators are then combined with a Wilkinson power combiner.

Although the extended resonance technique has been demonstrated at mm-wavelengths [6], for this design it was implemented at 7.53 GHz so that the difference in bond wire lengths between the two devices would have reduced impact. The IF signals were applied 180° out-of-phase using an external phase shifter (Mini-Circuits JSPHS-446). Lowpass filters permit the IF (400 MHz) to pass whilst the LO (7.5 GHz) lies in the stopband. This allows the gate of the oscillator to be loaded with a purely reactive load at the LO frequency to stimulate self-oscillation. The Wilkinson power combiner helps to isolate the devices from each other over a broad bandwidth and thus preserve input and output impedance levels.

III. Results

The devices were biased using external bias tees for maximum transconductance ($V_{DS} = 1.5$ V, $V_{GS} = -0.6$ V). Although the drain-source bias to the pHEMT's is delivered in union, the gate bias of each of the devices could be varied independently. When the oscillators were locked in anti-phase, a reduction in LO power (>10 dB) was experienced. It was found that tuning V_{GS} over ± 0.05 V for one of the devices was adequate in order to achieve a good balance and a reduction in LO power of -51.9 dBm (Fig. 2). The LO-RF isolation was estimated by pinching-off one of the devices and measuring the difference in LO power. An improvement in LO-RF isolation of 40.5 dB was found.

This circuit also relaxes the filtering requirements for generation of single sideband signals, which would be particularly difficult to achieve at mm-wave frequencies. Furthermore, it prevents the LO from saturating any power amplification stages before the sideband levels have reached suitable power levels for transmission.

Although the circuit has the added complexity of requiring two out-of-phase IF signals over the desired bandwidth, a 3 dB conversion gain bandwidth is obtained from $-70°$ to $+55°$ phase error as shown in Fig. 3. The phase was obtained using a vector voltage measurement on a transition analyzer

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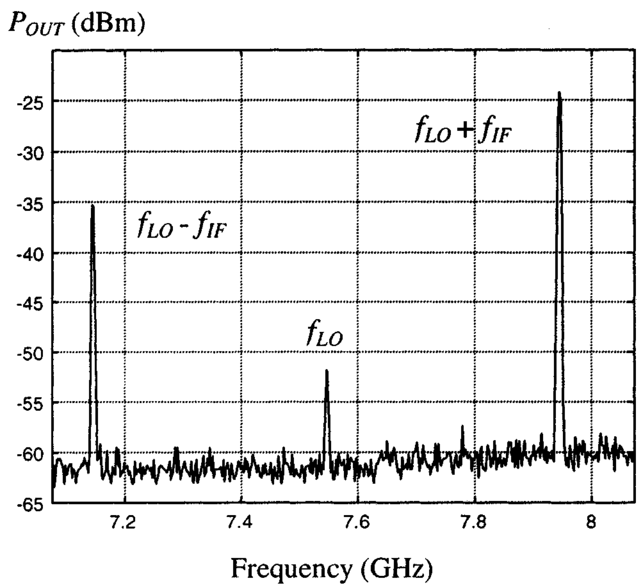

Fig. 2. Output from the balanced SOM.

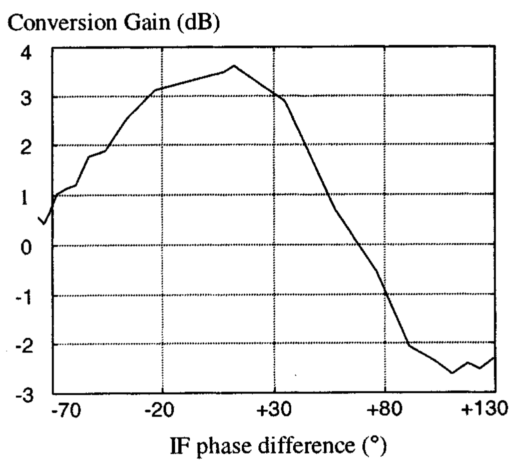

Fig. 3. Conversion gain dependence on IF phase imbalance.

(HP70820A). The circuit gave a 3.6 dB conversion gain for an IF of 400 MHz.

The balance in the oscillators occurs at $V_{DS} = 1.5$ V which is sufficiently away from the nonlinear knee region, where the conversion gain is improved, as shown in Fig. 4.

IV. Intermodulation Reduction

The two 400 MHz IF sources were phase locked together using the 10 MHz reference output from the crystal of one oscillator to stabilize the other. At the gate input of each oscillator there are two-tones (ω_1, ω_2) separated by 1 MHz:

$$V_{\text{in1}} = V_1 \cos(\omega_1 t) + V_2 \cos(\omega_2 t)$$
$$V_{\text{in2}} = V_1 \cos(\omega_1 t + \pi) + V_2 \cos(\omega_2 t + \pi).$$

Each self-oscillating mixer can be modeled using a power-series:

$$I_{\text{out}} = g_{m1} V_{\text{in}} + g_{m2} {V_{\text{in}}}^2 + g_{m3} {V_{\text{in}}}^3.$$

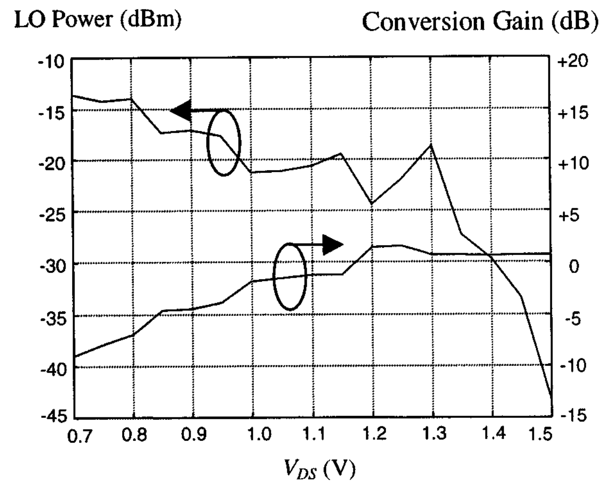

Fig. 4. Trade-off in LO power and conversion gain.

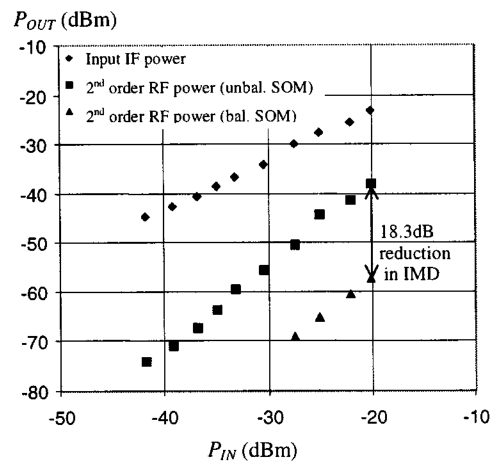

Fig. 5. Reduction in second order intermodulation.

The second-order current components at the outputs of the corresponding devices are

$$\begin{aligned} I_{\text{out1}} =& 0.5 g_{m2} \left(V_1^2 + V_2^2\right) + 0.5 g_{m2} V_1^2 \cos(2\omega_1 t) \\ &+ 0.5 g_{m2} V_2^2 \cos(2\omega_2 t) + 0.5 g_{m2} V_1 V_2 \\ &\cdot \cos(\omega_1 t + \omega_2 t) + 0.5 g_{m2} V_1 V_2 \cos(\omega_1 t - \omega_2 t) \end{aligned}$$

$$\begin{aligned} I_{\text{out2}} =& 0.5 g_{m2} \left(V_1^2 + V_2^2\right) + 0.5 g_{m2} V_1^2 \cos(2\omega_1 t + 2\pi) \\ &+ 0.5 g_{m2} V_2^2 \cos(2\omega_2 + 2\pi) + 0.5 g_{m2} V_1 V_2 \\ &\cdot \cos(\omega_1 t - \omega_2 t) + 0.5 g_{m2} V_1 V_2 \\ &\cdot \cos(\omega_1 t + \omega_2 t + 2\pi). \end{aligned}$$

However, with the balanced oscillators being out of phase, a further phase shift of 180° is applied to one of the mixed signals (I_{out}). Thus when I_{out1} and I_{out2} are added together by the Wilkinson power combiner, the second-order products are in anti-phase and cancel. Similar analysis for the odd-order products shows that they add in phase at the output.

The balanced SOM gave an 18.3 dB reduction in second-order intermodulation products and an extrapolated intercept point of +19.5 dBm compared to an unbalanced SOM (Fig. 5).

V. Conclusions

A balanced self-oscillating upconverter has been investigated which exhibits 3.6 dB of conversion gain, 40.5 dB of LO to RF isolation and 18.3 dB of second order intermodulation rejection. The circuit uses extended resonance to achieve a 180° phase balance between the two oscillators which are then combined in a Wilkinson coupler. Although these results have been obtained with a hybrid, this approach is fully compatible with MMIC technology.

Acknowledgment

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