Conversion Matrix and Gain of Self-Oscillating Mixers

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Abstract —**The conversion matrix of self-oscillating mixers is derived from the bias-, amplitude-, and frequency-dependent admittance of the active device together with its dynamic current–voltage characteristic. Components at the image frequency are also taken into account. With this matrix and the circuit admittances at the different frequencies involved, the conversion gain can be expressed. For better insight into the relevant mechanisms, the conversion gain is subdivided into the amplitude response of the self-excited oscillation to an input signal and the demodulation caused by the device internal rectification. The formalism is applied to a simplified model of an oscillating BARITT diode. The resulting analytical expressions allow a discussion of the influence of different device and circuit parameters as well as a qualitative and quantitative comparison with experimental results from a self-oscillating BARITT-diode mixer operating in the V band at 60 GHz.**

I. Introduction

THE principle and benefits of self-oscillating mixers have been described by several authors [1]–[3]. Practical microwave and millimeter-wave circuits with IMPATT diodes [1], Gunn devices [4], and especially BARITT diodes [1], [2], [5] working as local oscillators and, owing to their nonlinear behavior, as mixers as well have resulted in cheap and simple Doppler radar detectors or receiver front ends. Compared with IMPATT or Gunn devices, the good mixing properties of BARITT diodes, which lead to a substantial conversion gain and very low minimum detectable signal level [2], [5], favor the application of these devices in self-oscillating mixer circuits. New devices such as tunnel transit-time diodes [6], however, may also be well suited for such applications.

The conversion behavior of active BARITT diode mixers was first investigated theoretically by Vanoverschelde *et al.* [3] and then by Harth [7]. Both used a conversion matrix representation derived from an analytic impedance model of the device neglecting the components at the image frequency. This is not appropriate in most practical applications, e.g. Doppler radar systems, when the signal and the image frequency are relatively close to the frequency of the self-oscillation such that they are both within the bandwidth of the resonator circuit. Thus unrealistically high conversion gain and incorrect frequency dependence were calculated. The reason for this is that if a signal is injected into the oscillator circuit at one sideband, the amplitude saturation mechanism of the oscillator will produce an opposite sideband (i.e., a signal at the image frequency) such that the resulting amplitude modulation of the oscillator is much less than the phase modulation. The amplitude modulation only, however, is down-converted to the mixer output frequency owing to the device internal rectification.

In this paper a conversion matrix formulation is derived from a general linearized Taylor series for the amplitude-, frequency-, and dc-voltage-dependent device admittance in the RF and the bias circuit. This matrix (eq. (1)) contains all elements at the signal angular frequency ω_s, the image angular frequency $\omega_i = 2\omega_o - \omega_s$ (ω_o being the angular frequency of the free-running self-excited oscillation in the active mixer circuit), and the down-converted angular frequency $|\omega_d|$, where $\omega_d = \omega_s - \omega_o$ is the frequency difference between the signal and the oscillating mixer:

$$\begin{bmatrix} I_s \\ I_d \\ I_i^* \end{bmatrix} = \begin{bmatrix} Y_{ss} & Y_{sd} & Y_{si} \\ Y_{ds} & Y_{dd} & Y_{di} \\ Y_{is}^* & Y_{id}^* & Y_{ii}^* \end{bmatrix} \cdot \begin{bmatrix} U_s \\ U_d \\ U_i^* \end{bmatrix}. \quad (1)$$

Here I_s, I_d, I_i and U_s, U_d, U_i are the complex current amplitudes in the active device and the voltage amplitudes across the device at the frequencies ω_s, $|\omega_d|$, and ω_i, respectively; Y_{jk} are the conversion matrix elements; and the asterisk denotes conjugate complex quantities.

Together with the network properties of the surrounding mixer circuitry, the conversion behavior can be deduced from the conversion matrix. The treatment and the results are, however, rather complex and the final equations are difficult to analyze and discuss. The derivation of the conversion gain is, therefore, divided into two steps.

First, the amplitude modulation response of the oscillator due to the injected signal is calculated. This is discussed in terms of common disturbed oscillator theory (e.g. Kurokawa [8]). It can lead to high signal amplification at low oscillator amplitudes where the active device behaves rather linearly with weak amplitude saturation. The amplitude modulation is then down-converted into the bias circuit owing to the device internal rectification. The down-conversion is in general connected with signal losses, especially at low oscillator amplitudes, where the nonlinearities in the device are too small for efficient rectification. The total conversion gain is the product of the two processes.

For purposes of discussion and for a quantitative evaluation of the conversion behavior, simplified general expressions for the device admittance with quadratic amplitude saturation and for the dynamic bias current–voltage characteristic with quadratic rectification are used. Fitting the parameters of the admittance such that the calculated output power corresponds to the measurements allows a quantitative comparison between the theoretical treatment and experimental results for the conversion gain of V-band self-oscillating BARITT diode mixers [5].

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II. Conversion Matrix

In the following, complex phasor notation is used for all alternating quantities. At the oscillating angular frequency ω the complex current amplitude $I_1(t)$ is related to the voltage amplitude $U_1(t)$ across the active device by the device admittance

$$I_1(t) = Y(U_o, |U_1|, \omega) \cdot U_1(t) \tag{2}$$

which depends in general on the bias voltage U_o, on the absolute value of the oscillation amplitude U_1, and on the oscillation frequency ω. Herein the variations in time of I_1 and U_1, as well as of U_o and ω, are assumed to be slow compared with the oscillation period to allow a description by a slowly time-varying admittance.

On the other hand, the bias current I_o depends also on U_o, on $|U_1|$, and eventually on the real part of the oscillation frequency ω_r owing to the dynamic current–voltage characteristic:

$$I_o(t) = I_o(U_o, |U_1|, \omega_r). \tag{3}$$

All time-varying quantities can be expressed by their stationary value without an input signal and a time-varying disturbance:

$$\begin{aligned} I_1(t) &= I_{10} + \delta I_1(t) \\ U_1(t) &= U_{10} + \delta U_1(t) \\ |U_1(t)| &= U_{10} + \delta|U_1(t)| \\ I_o(t) &= I_{oo} + \delta I_o(t) \\ U_o(t) &= U_{oo} + \delta U_o(t) \\ \omega(t) &= \omega_o + \delta\omega(t) \\ \omega_r(t) &= \omega_o + \delta\omega_r(t). \end{aligned} \tag{4}$$

In the case of the self-oscillating mixer the disturbances are always small compared with the stationary values such that (2) and (3) may be developed into Taylor series up to linear terms in the disturbances only:

$$\delta I_1 = U_{10}\left[\frac{\partial Y}{\partial U_o}\delta U_o + \frac{\partial Y}{\partial |U_1|}\delta|U_1| + \frac{\partial Y}{\partial \omega}\cdot\delta\omega\right] + Y_o \cdot \delta U_1 \tag{5}$$

$$\delta I_o = \frac{\partial I_o}{\partial U_o}\cdot\delta U_o + \frac{\partial I_o}{\partial |U_1|}\cdot\delta|U_1| + \frac{\partial I_o}{\partial \omega_r}\cdot\delta\omega_r. \tag{6}$$

In (5), $Y_o = Y(U_{oo}, U_{10}, \omega_o)$ is the undisturbed device admittance at the oscillation frequency.

The disturbances of the oscillator amplitudes δI_1 and δU_1 consist of both sidebands at the signal frequency $\omega_s = \omega_o + \omega_d$ and at the image frequency $\omega_i = \omega_o - \omega_d$:

$$\delta I_1(t) = I_s e^{j\omega_d t} + I_i e^{-j\omega_d t} \tag{7a}$$

$$\delta U_1(t) = U_s e^{j\omega_d t} + U_i e^{-j\omega_d t}. \tag{7b}$$

In the bias circuit δI_o and δU_o can be expressed as

$$\delta I_o(t) = \frac{1}{2}\left(I_d e^{j\omega_d t} + I_d^* e^{-j\omega_d t}\right) \tag{8a}$$

$$\delta U_o(t) = \frac{1}{2}\left(U_d e^{j\omega_d t} + U_d^* e^{-j\omega_d t}\right). \tag{8b}$$

From (7b) the frequency deviation can be deduced:

$$\delta\omega = -j\frac{1}{U_{10}}\cdot\frac{dU_1}{dt} = \frac{\omega_d}{U_{10}}\left(U_s e^{j\omega_d t} - U_i e^{-j\omega_d t}\right) \tag{9}$$

with the real part

$$\delta\omega_r = \frac{\omega_d}{2\cdot U_{10}}\left[(U_s - U_i^*)e^{j\omega_d t} + (U_s^* - U_i)e^{-j\omega_d t}\right]. \tag{10}$$

The deviation of the absolute value of the amplitude is

$$\delta|U_1| = \frac{1}{2}\left[(U_s + U_i^*)e^{j\omega_d t} + (U_s^* + U_i)e^{-j\omega_d t}\right]. \tag{11}$$

Inserting (7), (8), (9), (10), and (11) into (5) and (6) and separately collecting the terms with $e^{j\omega_d t}$ and $e^{-j\omega_d t}$ leads to the elements of the conversion matrix in (1):

$$Y_{ss} = Y_o + \frac{U_{10}}{2}\cdot\frac{\partial Y}{\partial|U_1|} + \omega_d\frac{\partial Y}{\partial\omega} \tag{12a}$$

$$Y_{sd} = \frac{U_{10}}{2}\cdot\frac{\partial Y}{\partial U_o} \tag{12b}$$

$$Y_{si} = \frac{U_{10}}{2}\cdot\frac{\partial Y}{\partial|U_1|} \tag{12c}$$

$$Y_{ds} = \frac{\partial I_o}{\partial|U_1|} + \frac{\omega_d}{U_{10}}\cdot\frac{\partial I_o}{\partial\omega_r} \tag{12d}$$

$$Y_{dd} = \frac{\partial I_o}{\partial U_o} \tag{12e}$$

$$Y_{di} = \frac{\partial I_o}{\partial|U_1|} - \frac{\omega_d}{U_{10}}\cdot\frac{\partial I_o}{\partial\omega_r} \tag{12f}$$

$$Y_{is} = \frac{U_{10}}{2}\cdot\frac{\partial Y}{\partial|U_1|} \tag{12g}$$

$$Y_{id} = \frac{U_{10}}{2}\cdot\frac{\partial Y}{\partial U_o} \tag{12h}$$

$$Y_{ii} = Y_o + \frac{U_{10}}{2}\cdot\frac{\partial Y}{\partial|U_1|} - \omega_d\frac{\partial Y}{\partial\omega}. \tag{12i}$$

This is valid for all devices for which the admittance and the dynamic current–voltage characteristic can be expressed in the form of (2) and (3), respectively.

III. Conversion Gain

Fig. 1 shows the equivalent circuits of the self-oscillating mixer at the angular frequencies ω_s, ω_i, and ω_d, respectively. In these Y_s, Y_i, and Y_d represent the circuit admittances at the three frequencies as seen from the active device. The current source at the frequency ω_s with the amplitude i_s characterizes the RF input signal that is to be down-converted. At the active device there are, therefore, the complex voltage amplitudes

$$U_s = (i_s - I_s)/Y_s \tag{13a}$$

$$U_d = -I_d/Y_d \tag{13b}$$

$$U_i = -I_i/Y_i. \tag{13c}$$

These equations, combined with the conversion matrix (1) consisting of the elements in (12a)–(12i), determine all current and voltage amplitudes for a given input-signal amplitude i_s and, hence, the total conversion gain. However, for

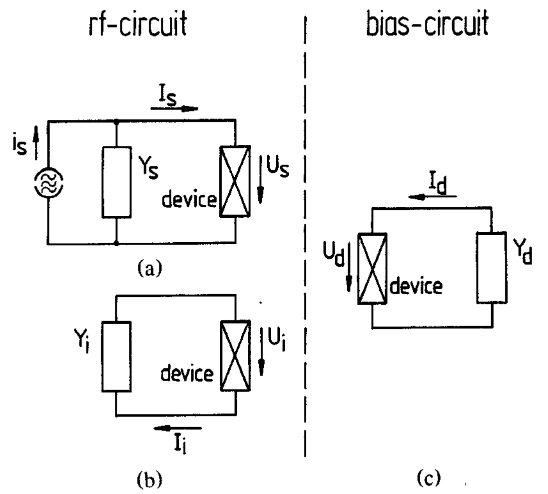

Fig. 1. Equivalent circuits of a self-oscillating mixer at (a) signal frequency ω_s, (b) image frequency ω_i, and (c) demodulated frequency ω_d.

better understanding it seems more appropriate to calculate first the amplitude modulation response of the oscillator due to the signal injection and then the demodulation of the amplitude fluctuations.

For this purpose (13b) can be inserted into (1) to transform the nine-element conversion matrix to a four-element matrix:

$$\begin{bmatrix} I_s \\ I_i^* \end{bmatrix} = \begin{bmatrix} Y_{ss}' & Y_{si}' \\ Y_{is}'^* & Y_{ii}'^* \end{bmatrix} \cdot \begin{bmatrix} U_s \\ U_i^* \end{bmatrix}. \tag{14}$$

The elements of this matrix become

$$Y_{ss}' = Y_{ss} - \frac{Y_{sd} Y_{ds}}{Y_{dd} + Y_d} \tag{15a}$$

$$Y_{si}' = Y_{si} - \frac{Y_{sd} Y_{di}}{Y_{dd} + Y_d} \tag{15b}$$

$$Y_{is}'^* = Y_{is}^* - \frac{Y_{id}^* Y_{ds}}{Y_{dd} + Y_d} \tag{15c}$$

$$Y_{ii}'^* = Y_{ii}^* - \frac{Y_{id}^* Y_{di}}{Y_{dd} + Y_d}. \tag{15d}$$

The maximum amplitude deviation,

$$\delta |U_1|_{\max} = |U_s + U_i^*| \tag{16}$$

is used to define an amplitude modulation gain:

$$g_{\mathrm{AM}} = \frac{1}{2} (\delta |U_1|_{\max})^2 \cdot G_L / P_s \tag{17}$$

where G_L is the load conductance of the RF circuit at the signal frequency and $P_s = i_s^2/(8G_L)$ is the signal input power. Using (13a), (13c), (14), and (16), the amplitude modulation gain (eq. (17)) is expressed as

$$g_{\mathrm{AM}} = \frac{4G_L^2 |Y_i^* + Y_{ii}'^* - Y_{is}'^*|^2}{|(Y_s + Y_{ss}')(Y_i^* + Y_{ii}'^*) - Y_{si}' Y_{is}'^*|^2}. \tag{18}$$

In the following the RF circuit admittance Y_c is replaced by that of a resonant circuit which is in parallel (or series) resonance with the active device at the oscillation frequency:

$$Y_c = -Y_o + \frac{dY_c}{d\omega}(\omega - \omega_o) \tag{19a}$$

i.e.,

$$Y_s = -Y_o + \omega_d \frac{dY_c}{d\omega} \tag{19b}$$

and

$$Y_i = -Y_o - \omega_d \frac{dY_c}{d\omega} \tag{19c}$$

such that the real part of the frequency dependence of the total admittance vanishes:

$$\mathrm{Re}\left\{ \frac{\partial Y}{\partial \omega} + \frac{dY_c}{d\omega} \right\} = 0. \tag{20}$$

In the bias circuit a real conductance $Y_d = G_B$ is used. In addition, the contribution of the frequency demodulation, expressed by the term $\omega_d / U_{10} \cdot \partial I_0 / \partial \omega_r$ in (12d) and (12f), is neglected, as it is in all practical cases much smaller than that of the amplitude demodulation. With this, (18) simplifies substantially and reaches a form similar to the amplitude noise formula derived by Edson [9] and by Kurokawa [8], where the amplitude response of the oscillator to a small signal near the oscillating frequency is also expressed:

$$g_{\mathrm{AM}} = \frac{1}{(s/2)^2 + [(\omega_d/\omega_o)Q_L]^2}. \tag{21}$$

Here s is the saturation factor for the real part of the device admittance:

$$s = \frac{U_{10} \cdot \mathrm{Re}\{\partial Y' / \partial |U_1|\}}{G_L} \tag{22}$$

including the feedback of the demodulated signal to the RF:

$$\frac{\partial Y'}{\partial |U_1|} = \frac{\partial Y}{\partial |U_1|} - \frac{\partial Y / \partial U_o \cdot \partial I_o / \partial |U_1|}{G_B + \partial I_o / \partial U_o}. \tag{23}$$

The quality factor Q_L of the oscillator is given by

$$Q_L = \frac{\omega_o \cdot \mathrm{Im}\{\partial Y/\partial \omega + dY_c/d\omega\}}{2G_L}. \tag{24}$$

In a similar manner a demodulation factor g_{dem} can be defined as

$$g_{\mathrm{dem}} = \frac{\frac{1}{2}|U_d^2| G_B}{\frac{1}{2}(\delta |U_1|_{\max})^2 G_L}. \tag{25}$$

This can be found applying I_d from (1) in (13b), making use of (16), and again neglecting $\partial I_o / \partial \omega_r$ in (12d) and (12f):

$$g_{\mathrm{dem}} = \frac{(\partial I_o / \partial |U_1|)^2}{G_L} \cdot \frac{G_B}{(G_B + \partial I_o / \partial U_o)^2}. \tag{26}$$

The demodulation factor contains the shift of the dynamic current–voltage characteristic with the amplitude $\partial I_o / \partial |U_1|$, which is responsible for the down-conversion, and the power matching in the bias circuit due to G_B.

The total conversion gain g_c of the self-oscillating mixer is the product of the amplitude modulation gain and the demodulation factor:

$$g_c = \frac{\frac{1}{2}|U_d|^2 G_B}{P_s} = g_{AM} \cdot g_{dem}. \tag{27}$$

IV. Device Model

To show the typical behavior of self-oscillating mixers, simplified expressions for the negative device RF conductance, $-G_D$, and the dynamic bias current–voltage characteristic are used which are in good qualitative agreement with experimental evidence from the behavior of, for example, oscillating BARITT diodes:

$$-G_D = G'\left[\Delta U_o - A(\Delta U_o)^2 - B|U_1|^2\right] - G_l \tag{28}$$

$$I_o = G_o\left[\Delta U_o + C|U_1|^2\right]. \tag{29}$$

ΔU_o is the bias voltage above reach-through, where the current flow sets in, G_l represents the RF losses in the diode, in which also circuit losses may be included; and G', G_o, A, B, and C are constants.

The negative RF conductance exhibits a maximum with respect to bias voltage, which leads to a maximum in the output power as found experimentally [2], [5], and quadratic amplitude saturation. The bias current increases linearly with dc voltage above reach-through and is shifted to higher current or lower voltage with increasing amplitude, also in a quadratic manner, which will at least be valid for low amplitudes, i.e., at low output power. With this the amplitude saturation factor (eqs. (22) and (23)) becomes

$$s = 2 \cdot G'B' \cdot U_{10}^2 / G_L \tag{30}$$

with

$$B' = B + C\frac{1 - 2A\Delta U_o}{1 + G_B/G_o}. \tag{31}$$

It increases with the square of the oscillation amplitude U_{10}. Consequently the amplitude modulation gain g_{AM} (eq. (21)) behaves at low output power and close to the oscillation frequency (small ω_d) as U_{10}^{-4}. The bandwidth of g_{AM}, however, is also proportional to U_{10}^2 and, therefore, decreases when the amplitude is reduced.

The demodulation factor g_{dem} (eq. (26)), on the other hand, is small at low amplitudes, where the rectification effect is weak, but it increases as U_{10}^2 with the oscillation amplitude. This, however, cannot compensate for the amplitude behavior of the amplitude modulation gain. Hence the total conversion gain g_c near the oscillation frequency,

$$g_c|_{\omega_d \approx 0} = \frac{1}{P_o} \cdot \frac{2 \cdot G_L^2 \cdot G_o \cdot C^2}{G'^2 \cdot B'^2} \cdot \frac{G_B/G_o}{(G_B/G_o + 1)^2} \tag{32}$$

increases at low power levels $P_o = U_{10}^2 G_L/2$ as $1/P_o$.

For comparison of (32) with experimental results, the parameters of (28) and (29) can be deduced from the measured oscillator power versus bias voltage (or current) characteristic at fixed matching and from the experimental static and dynamic bias current–voltage characteristics. This has been done for a V-band BARITT diode mixer [5] oscillating at 60 GHz with a maximum output power of $P_{max} = 1$ mW.

TABLE I

Symbol	Expression	Value
$\frac{G_l}{G'}$	$2\Delta U_{oo} - \frac{\Delta U_{oo}^2}{\Delta U_{o\,max}} - \frac{\Delta U_{o\,max}}{2}$	0.31 V
$\frac{G_L}{G'}$	$\frac{1}{2}\left[\frac{\Delta U_{o\,max}}{2} - \frac{G_l}{G'}\right]$	0.16 V
A	$\frac{1}{2\Delta U_{o\,max}}$	0.4 V^{-1}
$B\|U_1\|_{max}^2$	$\frac{1}{2}\left[\frac{\Delta U_{o\,max}}{2} - \frac{G_l}{G'}\right]$	0.16 V
$C\|U_1\|_{max}^2$	ΔU_{osd}	0.5 V
$P_o(\Delta U_o)$	$\frac{\Delta U_o - \Delta U_{oo} - A(\Delta U_o^2 - \Delta U_{oo}^2)}{B\|U_1\|_{max}^2} P_{max}$	$0 \cdots 1$ mW

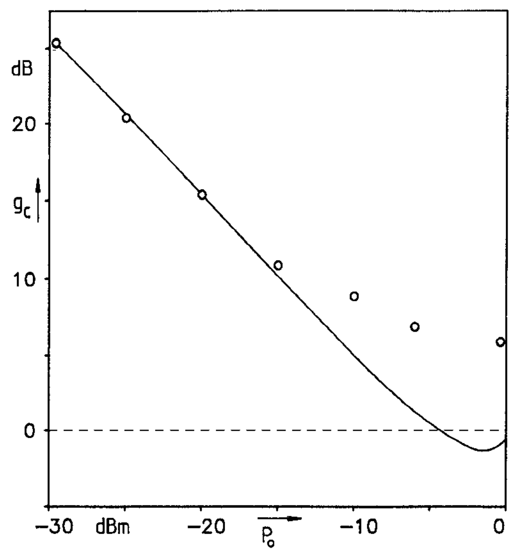

Fig. 2. Conversion gain g_c of self-oscillating V-band BARITT diode mixer versus output power P_o. Solid line calculated from eq. (32) with parameter values from Table I and $G_B = 10$ mmhos; circles measured with $\omega_d/2\pi = 1$ kHz.

The current–voltage characteristic of the nonoscillating diode had a reach-through voltage of $U_{th} = 21$ V and a slope of $G_o = 8$ mmhos. After tuning the mixer circuit to maximum output power the bias voltage difference $\Delta U_{o\,max} = 1.25$ V between the operating point and reach-through was measured as well as the voltage difference $\Delta U_{osd} = 0.5$ V at constant current between the static and the dynamic current–voltage characteristic (ΔU_{osd} is defined as positive when the dynamic bias voltage is lower than the static one). Finally the bias was reduced with fixed matching of the device ($G_L = -G_D =$ const.) until RF oscillation quenching. Here again the voltage difference $\Delta U_{oo} = 0.625$ V to reach-through was determined. From these data, which are fairly independent of the conversion gain measurements, all relevant quantities in (32) can be calculated. They are shown in Table I. Here $|U_1|_{max}^2$ is the square of the oscillation amplitude at maximum output power.

The conversion gain according to (32) using these values is depicted in Fig. 2 versus the oscillator output power P_o for a bias resistance $R_B = 1/G_B = 100\ \Omega$ and in Fig. 3 versus the bias resistance at a fixed output power of $P_o = -15$ dBm (solid lines). The circles correspond to the measured conver-

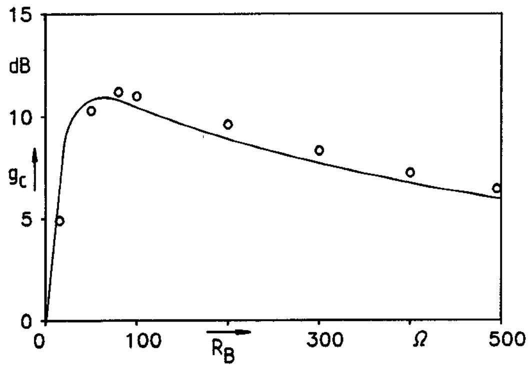

Fig. 3. Conversion gain g_c of self-oscillating V-band BARITT diode mixer versus bias resistance $R_B = 1/G_B$ with $P_o = -15$ dBm. Solid line calculated from eq. (32) with parameter values from Table I; circles measured with $\omega_d/2\pi = 1$ kHz.

sion gain [5] under equal conditions, i.e., with the device matched to maximum output power and then reducing the bias with constant RF load. The measurements were carried out with a signal frequency 1 kHz off the carrier well within the gain bandwidth.

V. Discussion

At low output powers, $P_o < -15$ dBm, the calculated and measured conversion gains agree rather well. Here it is to be expected that the quadratic amplitude dependence of admittance and dynamic current–voltage characteristic can be applied satisfactorily, leading to $g_{\mathrm{AM}} \propto 1/P_o^2$ and $g_{\mathrm{dem}} \propto P_o$ such that the total conversion gain g_c increases to low output power levels as $1/P_o$.

At higher power levels the slope of the conversion gain versus power decreases in the calculation because of less feedback from the bias circuit as the bias voltage approaches the value for maximum output power. In the measurements, the conversion gain is here even higher, which may be attributed to a less than quadratically increasing saturation of the actual negative device conductance at higher amplitudes.

With a signal frequency further off the carrier than the bandwidth of the amplitude modulation gain (e.g. more than 100 MHz) the behavior of the conversion gain may be quite the opposite. Then, according to (21), g_{AM} no longer depends on P_o but is determined by the quality factor of the circuit only. In this case the conversion gain increases with the output power owing to the increasing demodulation factor, as has also been found experimentally by Förg [10] and by Güttich [5].

Fig. 3 shows that in the theoretical curve as well as in the measurements the optimum matching of the demodulated signal is achieved at a bias resistance R_B of about 50 to 80 Ω, which is somewhat lower than the value corresponding to the slope of the current–voltage characteristic $1/G_o = 125$ Ω. This is due to the feedback of the demodulated signal to the RF circuit. At the output power maximum, where bias fluctuations do not lead to amplitude modulation the optimum matching is $R_B = 1/G_o$.

As to be seen from Table I the loss conductance G_l is a substantial part of the device negative conductance $-G_D$. This is not only important for the output power, which could be four times higher without losses, but also for the conversion gain (eq. (32)) because of the load conductance G_L, which would be twice as large for optimum power matching. Thus the conversion gain is equal in the output power maximum but it is higher at a distinct power. For example, at $P_o = -15$ dBm the calculated conversion gain becomes about 3 dB higher neglecting losses. This is less than a factor of 4 because the output power of -15 dBm is achieved at a lower bias voltage ΔU_o with $G_l = 0$, leading to stronger feedback and a larger B' according to (31).

On the other hand, a device (or circuit) with a maximum output power of only -10 dBm due to higher losses (all other parameters remaining constant) would need a load conductance that is lower by a factor of 0.32. The conversion gain of this device becomes $g_c = 18.2$ dB at an output power of -30 dBm, which is still quite high. Thus it is not only excellent devices that will lead to a substantial conversion gain at low output power.

Another interesting question is whether matching of the device to maximum output power is the optimum to achieve high conversion gain at low output power. According to (32), the load conductance G_L should be large and B' small. This can be achieved using a bias voltage according to the power maximum. In this point the highest small-signal conductance is reached, and $B' = B$ is rather small because there is no feedback. If the load conductance is increased close to the small-signal conductance of the device, the oscillating amplitude U_{10} becomes small, resulting also in low output power. At $P_o = -30$ dBm the load conductance is then higher by a factor of nearly 2 and B' is lower by a factor of 1.7 compared with maximum output power matching. A conversion gain of nearly 36 dB could thus be expected.

There are, however, some drawbacks to such an operating condition. One is that the device needs a high bias for low output power, leading to very low efficiency and also to high noise because of high current and temperature. On the other hand the condition of high load conductance is not easily achieved experimentally since a small load conductance can also lead to low output power. Finally, matching to the small-signal conductance is rather unstable. This is because even small changes in the load matching or the device parameters (for instance due to temperature fluctuation) affect the output performance and by this the conversion gain significantly or interrupt the oscillation. Thus matching the device to the output power maximum will for most applications (e.g. simple and robust Doppler sensors) be more practical.

For a first estimation of whether an active device is able to act as a self-oscillating mixer with sufficient conversion gain, (32) can further be simplified in that in the output power maximum $G_L/G' = B \cdot |U_1|^2_{\max}$ holds as well as $C|U_1|^2_{\max} = \Delta U_{osd}$ and a bias matching $G_B = G_o$ is used:

$$g_c P_o = \frac{B^2}{B'^2} \cdot \frac{1}{2} G_o \Delta U_{osd}^2. \tag{33}$$

Neglecting for the present the influence of the feedback from the bias circuit to the RF, which is included in the B/B' ratio, the right-hand side of (33) is only determined by the slope G_o of the bias current–voltage characteristic and by the shift ΔU_{osd} of the dynamic current–voltage characteristic in the output power maximum with respect to the static one. For the experimentally investigated BARITT diode, $G_o \Delta U_{osd}^2/2 = 1$ mW, leading to a conversion gain of $g_c = 30$

dB at an output power of $P_o = -30$ dBm when the feedback is ignored ($B/B' = 1$). With feedback the conversion gain will be lower, but in general by not more than 10 dB. Thus a device with a differential dc resistance of the order of 100 Ω should exhibit a dynamic shift of the current–voltage curve by a few tenths of a volt to achieve good conversion gain.

VI. Conclusion

In contrast to the formalism of Vanoverschelde *et al.* [3] and Harth [7], the conversion matrix of self-oscillating mixers was derived not from a time-varying admittance of the active device but from its amplitude and bias voltage dependence in conjunction with the dynamic current–voltage characteristic and was expanded by the components at the image frequency. This leads to a lower, more realistic conversion gain that can be brought into a form in terms of common disturbed oscillator theory to give more insight into the conversion gain mechanism.

A simplified admittance model for the BARITT diode was used to discuss the influence of oscillator output power, frequency set-off, bias circuit matching, device and circuit losses, and oscillator matching at the RF. It can be shown that the conversion gain increases at low oscillator power P_o with fixed matching as $1/P_o$. Device losses reduce the conversion gain by a factor smaller than the reduction in the maximum output power. Matching at the RF to optimum power output seems to be a good compromise with respect to conversion gain, efficiency, stability, and noise.

The parameters in the analytic expression for the conversion gain derived from the simplified device model could be determined experimentally by simple power, dc current, and voltage measurements. With these, rather good agreement was achieved between calculated and measured conversion gain quantitatively at low oscillator power and qualitatively near the power maximum as well.

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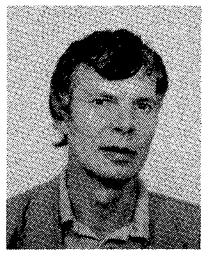

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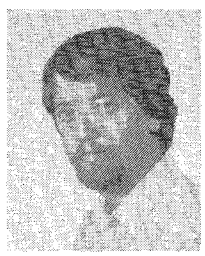

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