Simulation & Modelling Of A Heterojunction Bipolar Transistor Mixer

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Abstract

The first low power HBT mixer intended for use in mobile communications equipment is described. Measured conversion gain and intermodulation performance superior to active MESFET mixers is reported. An HBT model developed for mixer simulation is presented and accompanying predictions are given along with an analytical expression for predicting conversion gain.

Introduction

The intermodulation performance and high gain of the HBT have been demonstrated in power amplifier applications [1]. The intermodulation performance of a receiver is usually determined by the mixer stage, thus high IP_3 is desirable. The HBT mixer offers superior intermodulation and gain performance with respect to a similar active MESFET mixer [2]. The HBT mixer evaluated in this work is intended for use in future generation personal communications receivers, specified by the Groupe Special Mobile (GSM) standard. The allocated signal frequency band is UHF. Low power consumption is essential for mobile communications circuits. The HBT mixer offers a higher OIP_3-d.c ratio than an equivalent MESFET mixer.

HBT Non-linear Model

The HBT non-linear model is developed from a small signal equivalent circuit [1]. The HBT's equivalent circuit (fig 1) represents the device's small signal performance at X-band, thus, considerable simplification is possible at 950MHz. The model has series elements in its branches and it's overall form approaches a π circuit. To derive a tractable expression for conversion gain (*Gc*), only the signal, image and intermediate frequencies are considered. It is assumed that all other small signal mixing frequencies are terminated reactively. Given this assumption and the UHF signals specified, the model of fig 1 may be simplified by omitting all of the capacitors connected by dashed lines in fig 1. Further simplifications can be obtained by converting the current controlled current source (CCCS) into two voltage controlled current sources (VCCS). The dependence of the base emitter VCCS may be modelled by combining its current contribution with the base emitter resistance to form a combined resistor, R_j.

$$R_j = R_J + \alpha \qquad (1)$$

where R_J = junction resistance & α = common base current gain

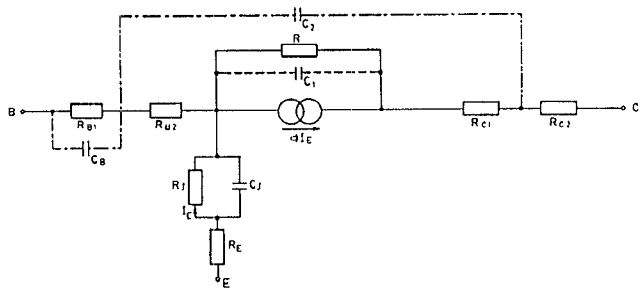

Fig 1: Small signal equavilent circuit of an HBT

The model is further simplified by representing the feedback resistor R in fig 1 as two resistors and two VCCS as shown in fig 2. The collector base resistance R, is of the order of MΩ [1], therefore, the current introduced by the VCCS (shown dashed in fig 2) is ignored because it is small compared to currents which exist in the base and collector circuits. In the base circuit, R, shown dashed in fig 2, is shunted by R_j, which is typically very much smaller than R. Hence, the parallel combination may be approximated by R_j. The final π model is given in fig 3. All components connected by dashed lines are omitted.

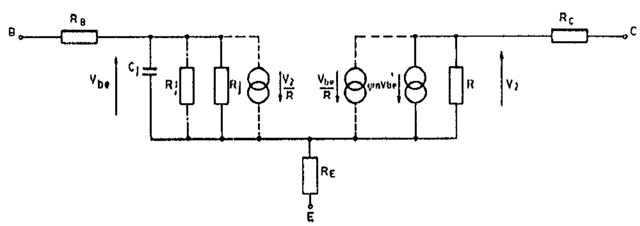

Fig 2: Simplified large signal model of an HBT

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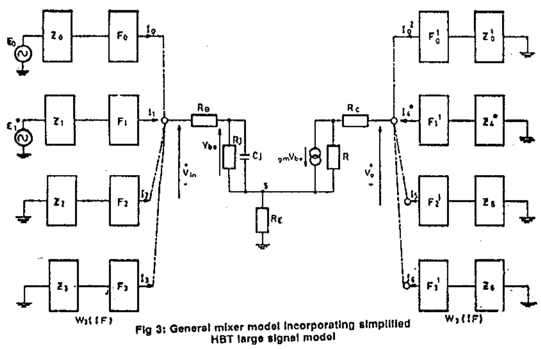

Fig 3: General mixer model incorporating simplified HBT large signal model

Model Parameter Extraction

The small signal S-parameters of the HBT used in this experiment were measured at varying bias levels over the frequency range 300kHz-3GHz. The bias levels were chosen to represent the voltage swing of the LO, which is assumed to be a sinusoid. The resultant S-parameters were de-embedded and the model of fig 1 was then fitted to each set of bias dependant data using an optimisation routine. Each parameter of the model was then plotted against bias to determine the degree of non-linearity. The dominant non-linearity was found to be the junction resistance; however the junction capacitance and collector base capacitance were also found to be non-linear. The transconductance of the π model was measured under d.c conditions by plotting I_C v V_{BE}. At frequencies less than 100kHz, the resistor R (see fig 1) is equal to reciprocal of the common base output conductance (h_{ob}), thus, using the two port relation:

$$h_{ob}=h_{oe}\cdot(1+\beta) \qquad (2)$$

where h_{oe} = common collector output resistance,
β = common emitter current gain

The resistor R can be determined from I_C v V_{CE} characteristics measured on a semiconductor parameter analyser and eqn 2.

Mixer Conversion Gain Expression

Using the non-linear model developed in the previous section, an expression for conversion gain can be derived. The HBT mixer model is incorporated into the model suggested by Pucel [3] (see fig 3). It is algebraically convenient to analyse this model using Z-parameter techniques because there are several impedances connected in series. A Norton source transformation is readily applied to the CCCS. It is assumed the signal voltage is considerably smaller than the local oscillator, so that the transconductance can be expanded into a Fourier series about the LO waveform. Only the fundamental (g_o) and first harmonic (g_1) of the Fourier series representation of the transconductance are required to generate image, signal and intermediate frequencies in the output current. Applying two port analysis techniques results in eqn 3,

$$\begin{bmatrix} E_1 \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} Z11' & 0 & 0 & Z14' & 0 & 0 \\ 0 & Z22 & 0 & 0 & Z25 & 0 \\ 0 & 0 & Z33 & 0 & 0 & Z36 \\ Z41' & 0 & Z43 & Z44' & 0 & 0 \\ 0 & Z52 & Z53 & 0 & Z55 & 0 \\ Z61' & Z62 & Z63 & 0 & 0 & Z66 \end{bmatrix} \times \begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \\ I_5 \\ I_6 \end{bmatrix} \qquad (3)$$

where the driving point impedances Znn (n=1,2,3,4,5,6) are the sum of the external embedding impedances (fig 3) Zm (m=1,2,3,4,5,6) and the device driving point impedances Zpp' (p=1,2,3,4,5,6).

Each Z-parameter of eqn 3 has been derived in terms of the equivalent circuit model, see Appendix I. Maximum conversion gain is obtained if both the signal and intermediate frequency ports are conjugately matched. Under this condition the maximum conversion gain is given by eqn 4.

$$G_{av}=\frac{|I_6|^2\cdot R_L}{|E_1|^2\cdot R_G} \qquad (4)$$

Where R_L = real part of load impedance, R_G = real part of generator impedance, I_6 = I.F output current, E_1 = Signal input voltage

The expression for the general case of a mixer described by fig 3 is determined by applying Cramers rule to solve the ratio of intermediate frequency current I_6 to signal voltage E_1 and is given in eqn 5 below.

$$\frac{I_6}{E_1}=\frac{Z22\cdot Z33\cdot Z44\cdot Z55\cdot Z61-Z25\cdot Z33\cdot Z44\cdot Z52\cdot Z61}{\begin{bmatrix} Z11\cdot Z22\cdot Z33\cdot Z44\cdot Z55\cdot Z66 & -Z11\cdot Z22\cdot Z36\cdot Z44\cdot Z55\cdot Z63 \\ +Z11\cdot Z25\cdot Z36\cdot Z44\cdot Z52\cdot Z63 & -Z11\cdot Z25\cdot Z36\cdot Z44\cdot Z53\cdot Z62 \\ -Z11\cdot Z25\cdot Z33\cdot Z44\cdot Z52\cdot Z66 & -Z14\cdot Z25\cdot Z36\cdot Z41\cdot Z52\cdot Z63 \\ +Z14\cdot Z25\cdot Z36\cdot Z41\cdot Z53\cdot Z62 & +Z14\cdot Z25\cdot Z36\cdot Z43\cdot Z52\cdot Z61 \\ +Z14\cdot Z25\cdot Z33\cdot Z41\cdot Z52\cdot Z66 & +Z14\cdot Z22\cdot Z36\cdot Z41\cdot Z55\cdot Z63 \\ -Z14\cdot Z22\cdot Z36\cdot Z43\cdot Z55\cdot Z61 & -Z14\cdot Z22\cdot Z33\cdot Z41\cdot Z55\cdot Z66 \end{bmatrix}} \qquad (5)$$

To simplify the analysis the following assumptions are made:

i) Z44'= Z55'= Z66'= R as R_E << R >> R_j (device driving point impedances)

ii) $jwC_j \approx 0$ @ 950MHz

Applying these conditions to eqn 5 results in eqn 6:

$$\frac{I_6}{V_1}=\frac{-g_1\cdot R_j\cdot R(Z22+g_0\cdot R_E\cdot R)}{\begin{bmatrix} Z11\cdot Z22\cdot Z66 & +Z11\cdot R_E\cdot g_0\cdot R_j\cdot R & +\frac{Z11\cdot R_E^2\cdot g_0^2\cdot R_j^2\cdot R}{Z33} \\ -\frac{Z11\cdot R_E^2\cdot g_1^2\cdot R_j^2\cdot R}{Z33} & +Z66\cdot Z11\cdot R_E\cdot g_0\cdot R & +\frac{R_E^3\cdot g_0^3\cdot R_j^3\cdot R}{Z33} \\ -\frac{R_E^3\cdot g_1^2\cdot g_0\cdot R_j^3\cdot R}{Z33} & +R_E^2\cdot g_0^2\cdot R_j^2\cdot Z66 & +\frac{Z22\cdot R_E^2\cdot g_0^2\cdot R_j^2\cdot R}{Z33} \\ -\frac{Z22\cdot g_1^2\cdot R_j^2\cdot R_E^2\cdot R}{Z33} & +Z22\cdot R_E\cdot g_0\cdot R_j\cdot Z66 & -\frac{R_E^3\cdot R_j^3\cdot g_1^2\cdot g_0\cdot R}{Z33} \end{bmatrix}} \qquad (6)$$

The device driving point impedances, Z11', Z22', Z33' are equal since $jwC_j \approx 0$. Z2, the image embedding impedance is much greater than the signal embedding impedance Z1, therefore Z22 >> Z11. Further, assuming Z22 >> $g_0.R_E.R_j$ results in eqn 7, a simple expression for determining *Gc* of an HBT mixer.

$$Gc = \frac{g_1^2 \cdot R_j^2 \cdot R_L}{4 \cdot R_G} \qquad (7)$$

Equation 7 can be used to predict the conversion gain of an HBT mixer. With $g_1 = 0.33$, $R_j = 0.6$, R_L = 1MΩ, R_G = 50Ω, eqn 6 predicts *Gc* = 23dB. Approximating eqn 6 by eqn 7 results in a ±4% error with Z22 = 500R.

Computer Simulation Of HBT Mixer

The conversion gain expression (eqn 5) for the general case of a single ended mixer, has been incorporated into a computer program to allow the user to access the various affects of terminating impedances and device parameters. A plot of HBT conversion gain versus signal terminating impedance and intermediate frequency terminating impedance is shown in fig 4. The computer program allows the use of more detailed HBT models to describe mixer performance. Using the program, the conversion gain was found to be most sensitive to IF output terminations, signal terminations and transconductance, as is the case for a MESFET [3]. Since both the output resistance and transconductance of an HBT are greater than the same parameters in a MESFET, the conversion gain of an HBT mixer is predicted to be superior to a MESFET device.

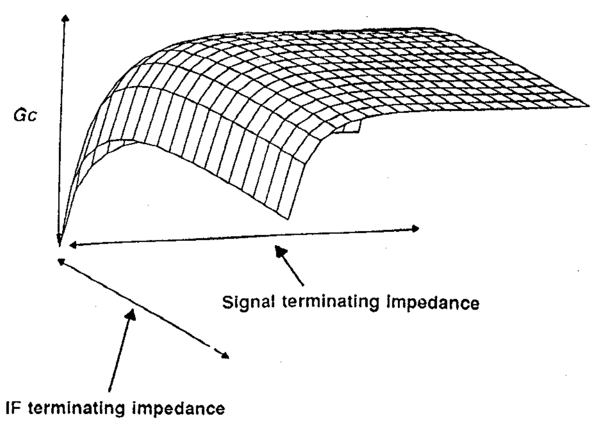

Fig 4: Computer simulation of Gc v IF & signal terminating impedances

Measurements

The measurement apparatus used in this experiment is shown in fig 5. The HBTs were bonded to Alumina carriers and mounted in a brass test jig. The local oscillator and signal ports were independently tuned using two slug tuners. The IF port was tuned using a lumped element circuit. A 3dB hybrid was used to couple LO and signal. Intermodulation was measured in the usual way by applying two closely spaced tones through a 3dB hybrid.

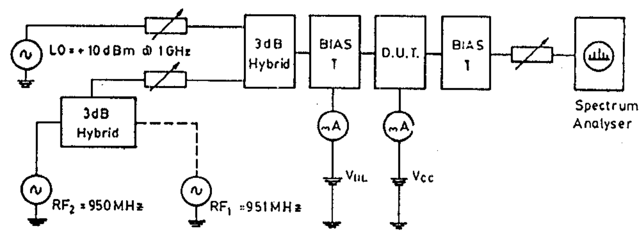

Fig 5: HBT mixer measurement

Results

The conversion gain was measured to be 21.5dBm with all ports tuned and with, LO=+1dBm,I_C=10mA and V_{CE}=5V. This figure is superior to the best reported figures for active MESFET mixers [2]. The third order intercept point, referenced to the output, was found to be +19dBm (fig 6) without the IF port tuned. Conversion gain in this case was +14.5dB. The OIP_3-d.c power ratio is 2.1, which is superior to the typical value of a MESFET of 0.5 [2], thus emphasizing the HBT's low d.c power operation capabilities and its suitability for use in personal communications receivers. The 1dB compression point of the mixer, referenced to the output, is +5dBm (fig 6). This figure is 4dB above the LO drive level, thus, the HBT mixer's signal handling ability is better than an active MESFET mixer [2].

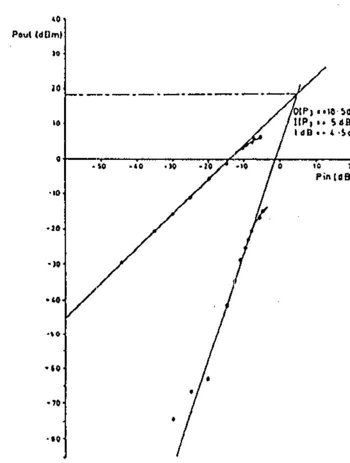

Fig 6: Intermodulation performance of an HBT mixer

Measured v Modelled

The predicted conversion gain is 1.5db greater than that measured. The error is caused by the simplified model, the number of non-linearities considered and the limited number of frequencies considered in the analysis. However eqn 6 offers a useful prediction of HBT mixer performance.

Mobile Communications Applications

The HBT offers the following advantages for mixers intended for use in mobile communications receivers compared with MESFETs:

a) Superior *Gc* (Reduced number of IF stages)

b) Increased OIP_3-d.c ratio (better adjacent channel interference properties).

c) Low d.c power operation (Increased battery life).

d) Higher 1dB Compression (Better signal handling capabilities)

Conclusions

The HBT offers superior conversion gain and intermodulation for a lower d.c power consumption than active MESFET mixers. The HBT offers several advantages for personal communications receivers including, longer battery life, reduced IF circuit complexity and better adjacent channel interference properties. Both MESFET and HBT mixer conversion gain is a strong function of output resistance and transconductance, hence, the HBT offers better conversion gain. An analytical expression has been derived which enables prediction of the conversion gain of an HBT mixer to be made. A computer program has been implemented which enables the user to assess the effect of model parameters and embedding impedances on conversion gain.

Appendix I

Derivation of the Z-parameter expressions in terms of the mixer model presented in fig 3.

$$Zkk(w_k)=R_B+R_E+\frac{R_J}{1+j\cdot\omega_k\cdot C_J\cdot R_J} \qquad (k=1,2,3) \qquad (I.1)$$

$$Zkk(\omega_k)=R_E+R_C+R \qquad (k=4,5,6) \qquad (I.2)$$

$$Z14=Z25=Z36=R_E \qquad (I.3)$$

$$Z41=R_E-\frac{g_0\cdot R_J\cdot R}{1+j\cdot\omega_1\cdot C_J\cdot R_J} \qquad (I.4)$$

$$Z52=R_E-\frac{g_0\cdot R_J\cdot R}{1+j\cdot\omega_2\cdot C_J\cdot R_J} \qquad (I.5)$$

$$Z63=R_E-\frac{g_0\cdot R_J\cdot R}{1+J\cdot\omega_3\cdot C_J\cdot R_J} \qquad (I.6)$$

$$Z61=\frac{-g_1\cdot R_J\cdot R}{1+j\cdot\omega_1\cdot C_J\cdot R_J} \qquad (I.7)$$

$$Z62=\frac{-g_1\cdot R_J\cdot R}{1+j\cdot\omega_2\cdot C_J\cdot R_J} \qquad (I.8)$$

$$Z43=Z53=\frac{-g_1\cdot R_J\cdot R}{1+j\cdot\omega_3\cdot C_J\cdot R_J} \qquad (I.9)$$

where R_E = emitter resistance, R_C = collector resistance, R_B = base resistance, C_j = junction capacitance, R_j = junction resistance, g_0 = fundamental of Fourier series of transconductance, g_1 = first harmonic of Fourier series of transconductance.

References

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