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Abstract—In this paper, the nonlinear model of the GaAs HEMT active Gilbert cell mixer is investigated. Based on the model, the conversion gain expression of active Gilbert cell mixers is derived theoretically by using conversion matrix analysis method. The expression is verified by harmonic balance simulation with Angelov HEMT model in Agilent Advanced Design System (ADS) and by chip measurement results.

Index Terms—Active Gilbert cell mixers, Conversion gain, Conversion matrix analysis, GaAs HEMT.

I. INTRODUCTION

In radar and wireless communication systems, the mixer plays a very important role in the way of providing frequency conversion between different frequency bands. The active Gilbert cell mixer [1] shown in Fig. 1 (a) is now widely used in this application. In the mixer, T1-T4 compose a switching quad, and T5-T6 represents the transconductance stage. The main virtues for this topology are its high conversion gain in a broad frequency range and excellent port-to-port isolation. In order to investigate the performance of the active mixer for wideband application, a theoretical analysis on the main bandwidth limitations in the Gilbert cell topology is required. When the large LO signals are fed to the switching quad of the Gilbert cell, the switching quad can be represented by a linear time-varying circuit. Therefore, a traditional linear time-invariant analysis on such circuit becomes invalid. Instead, conversion matrix analysis [2] is applied in order to obtain an analytical expression for the conversion gain as a function of the RF input frequency and IF output frequency. This paper describes a theoretical analysis of the conversion gain of GaAs HEMT active mixers based on the Gilbert cell topology. An analytical expression is derived which is capable of predicting the main bandwidth limitations for active mixers performing frequency conversion of wideband signals. The analysis is verified by harmonic balance simulations in Agilent Advanced Design System (ADS) using the nonlinear HEMT Angelov model. This paper presents an extension of the
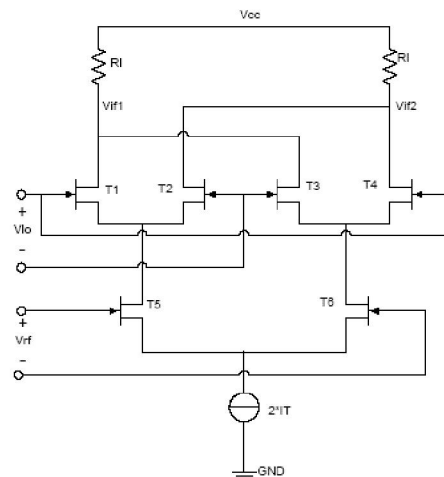

Fig. 1. The GaAs HEMT Active Gilbert cell mixer

analysis by the authors on HBT device to include HEMT device [3, 4].

II. CONVERSION MATRIX ANALYSIS

Conversion matrix analysis is a powerful tool for solving problems where a nonlinear device is driven by a single large sinusoidal signal and a much smaller signal. It is a suitable technique for analyzing and designing mixers. First, nonlinear devices driven by large signals are analyzed by the harmonic-balance method. The nonlinear elements in the device's equivalent circuit are then linearized to create small-signal linear time-varying elements. Finally, a small–signal analysis can be performed on the time-varying circuit.

Fig. 2 shows the time-varying small-signal equivalent circuit model of the half Gilbert cell mixer. The switching quad is composed of time-varying elements, and the transconductance stage is composed of I_s and Z_{ds}. The switching quad driven by the large LO signal is suitable for the conversion matrix analysis, and the transconductance stage driven by small RF
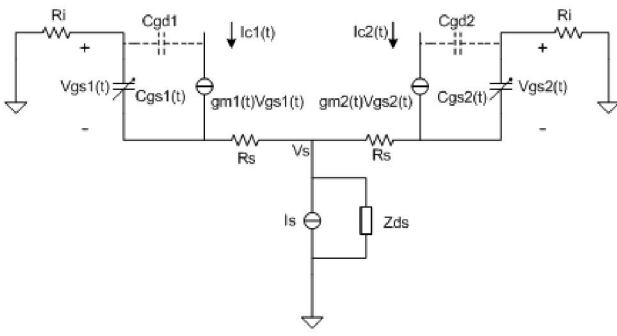
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Fig. 2. The small-signal linear time-varying equivalent circuit model for GaAs HMET active Gilbert cell mixers

signal is suitable for the time-invariant linear analysis.

A. Switch quad analysis

According to the behavior of the mixer, the time-varying elements in the Gilbert cell mixer excited by the small RF signal result in mixing frequencies. These frequencies can be represented as

$$\omega_n = \omega_{if} + n\omega_{LO} \qquad n = 0, \pm1, \pm2, \pm3,...... \qquad (1)$$

where $\omega_{if} = |\omega_{LO} - \omega_{rf}|$ is the IF angular frequency, ω_{rf} and ω_{LO} are the RF and LO angular frequency respectively. The frequencies interested can be expressed by

$$\omega_n = \begin{cases} \omega_{if} & for \quad n = 0 \\ \omega_{rf} & for \quad n = 1 \\ 2\omega_{if} - \omega_{rf} & for \quad n = -1 \end{cases} \qquad (2)$$

Because the switch quad is excited by the large LO signal, the time-varying nonlinear transconductance of transistors in the switch quad can be expressed by

$$g_m(t) = \begin{cases} G_{max} & for |\omega_{LO}t - 2n\pi| < (\pi - \alpha)/2 \\ 0 & else \end{cases} \qquad (3)$$

where the G_{max} is the maximum transconductance of transistors in the switching quad and α is the conduction angle in one period. In order to perform the conversion matrix calculation, the Fourier coefficient of the transconductance should be derived to compose the related conversion matrix. These are given by

$$G_n = \begin{cases} \dfrac{\alpha}{\pi}G_{max} & for \quad n = 0 \\ \dfrac{\sin\alpha}{\pi}G_{max} & for \quad n = \pm1 \\ \dfrac{\sin 2\alpha}{2\pi}G_{max} & for \quad n = \pm2 \end{cases} \qquad (4)$$

Normally, α is equal to $\pi/2$ because the duty cycle is considered ideal here. So, we can obtain the conversion matrix of the transconductance given by

$$\underline{G}_m = \begin{bmatrix} G_0 & G_{-1} & G_{-2} \\ G_1 & G_0 & G_{-1} \\ G_2 & G_1 & G_0 \end{bmatrix} = \begin{bmatrix} \dfrac{G_{max}}{2} & \dfrac{G_{max}}{\pi} & 0 \\ \dfrac{G_{max}}{\pi} & \dfrac{G_{max}}{2} & \dfrac{G_{max}}{\pi} \\ 0 & \dfrac{G_{max}}{\pi} & \dfrac{G_{max}}{2} \end{bmatrix} \qquad (5)$$

In order to analyze the time-varying small-signal equivalent circuit of the two switching stage transistors shown in Fig. 2, Kirchoff's law can be utilized here. The gate-source voltage vector $\underline{V}_{gs1}$ and $\underline{V}_{gs2}$ is expressed in terms of the voltage vector $\underline{V}_s$ at the common source point as

$$\underline{V}_{gs1} = -[\underline{I} + \underline{G}_{m1}\underline{R}_s + j\Omega\underline{C}_{gs1}(\underline{R}_i + \underline{R}_s)]^{-1}\underline{V}_s \qquad (6)$$

$$\underline{V}_{gs2} = -[\underline{I} + \underline{G}_{m2}\underline{R}_s + j\Omega\underline{C}_{gs2}(\underline{R}_i + \underline{R}_s)]^{-1}\underline{V}_s \qquad (7)$$

where $\underline{I}$ is the identity matrix, $\underline{G}_{m1}$ is the conversion matrix for $g_{m1}(t)$, $\underline{G}_{m2}$ is the conversion matrix for $g_{m1}(t)$, $\underline{C}_{gs1}$ is the conversion matrix for $C_{gs1}(t)$, $\underline{C}_{gs2}$ is the conversion matrix for $C_{gs2}(t)$, Ω is the diagonal matrix comprised by mixing frequency, $\underline{R}_s$ is the diagonal matrix with elements R_s, and $\underline{R}_i$ is the diagonal matrix with elements R_i. The drain currents of transistors in switch quad can be expressed by the transconductance matrix and the voltage vector of $V_{gs}(t)$:

$$\underline{I}_{c1} = \underline{G}_{m1}\underline{V}_{gs1} \qquad (8)$$

$$\underline{I}_{c2} = \underline{G}_{m2}\underline{V}_{gs2} \qquad (9)$$

Considering the common source point using Kirchoff's current law, the tail current $\underline{I}_s$ can be expressed by

$$\underline{I}_s = \underline{I}_{c1} + \underline{I}_{c2} + j\Omega\underline{C}_{gs1}\underline{V}_{gs1} + j\Omega\underline{C}_{gs2}\underline{V}_{gs2} - \dfrac{\underline{V}_s}{\underline{Z}_{ds}} \qquad (10)$$

where $\underline{I}_s$ is the small-signal excitation vector. There is only one non-zero element in the current vector, which comes from the transconductance stage at RF frequency. Because the transistors at the transconductance stage are considered as linear time-invariant model, the currents at other harmonic frequencies are neglected.

Substituting (6), (7), (8) and (9) into (10), the current $\underline{I}_s$ can be expressed in terms of $\underline{V}_s$ as

$$\underline{I}_s = -[(\underline{G}_{m1} + j\Omega\underline{C}_{gs1})[\underline{I} + \underline{G}_{m1}\underline{R}_s + j\Omega\underline{C}_{gs1}(\underline{R}_i + \underline{R}_s)]^{-1} + \qquad (11)$$
$$(\underline{G}_{m2} + j\Omega\underline{C}_{gs2})[\underline{I} + \underline{G}_{m2}\underline{R}_s + j\Omega\underline{C}_{gs2}(\underline{R}_i + \underline{R}_s)]^{-1} + \underline{Z}_{ds}^{-1}]\bullet\underline{V}_s$$

The conversion gain of the switch quad (G_{sw}) is defined as the ratio between the differential IF output voltage and the RF input current generated by the transconductance stage, which can be expressed by

$$\underline{G}_{sw} = \dfrac{\underline{V}_{if2} - \underline{V}_{if1}}{\underline{I}_s} = \dfrac{(\underline{I}_{c2} - \underline{I}_{c1})}{\underline{I}_s}\underline{R}_L \qquad (12)$$

where $\underline{R}_L$ is the diagonal matrix with the load resistor. Combining (6), (7), (8), (9), (11) and (12) and setting R_s to be zero, the G_{sw} is derived as the below expression:

$$G_{sw} = \dfrac{2}{\pi}\dfrac{A}{B + jC} \qquad (13)$$
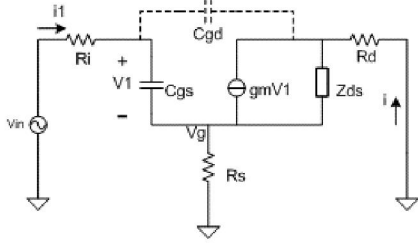
where $\quad A = R_{ds}G_{max}R_L$

Fig. 3. The small-signal equivalent circuit for GaAs HMET

$$B = 1 + R_{ds}G_{max} - R_{ds}R_iC_{gs0}C_{ds0}\omega_{rf}^2$$

$$C = \omega_{rf}C_{gs0}R_i + \omega_{rf}C_{ds0}R_{ds} + 2\omega_{rf}C_{gs0}R_{ds}$$

B. Transconductance stage analysis

Because the transconductance stage is excited by a small RF signal, the linear time-invariant analysis method can be utilized here. Fig.3 shows the simple small-signal GaAs HEMT model used in the analysis. The C_{gd} is not included to simplify the analysis process. The conversion gain of the transconductance stage (G_{tr}) is the ratio between the RF current used to bias the switching quad and the input RF voltage. It is derived as:

$$G_{tr} = \frac{g_{m\_tr}Z_{ds} - j\omega_{rf}C_{gs\_tr}R_{s\_tr}}{R_{s\_tr}(g_{m\_tr}Z_{ds} - j\omega_{rf}C_{gs\_tr}R_{s\_tr}) + [j\omega_{rf}C_{gs\_tr}(R_{i\_tr} + R_{s\_tr}) + 1](R_{\_tr} + Z_{ds})} \quad (14)$$

where Z_{ds} is composed of C_{ds} and R_{ds} in parallel and the parameters denoted by "_tr" are the parasitic components in the transconductance stage. The overall conversion gain of the mixer (G_{cnv}) is the product of G_{sw} and G_{tr}, which is given as

$$G_{cnv} = \frac{V_{if1} - V_{if2}}{V_{rf}} = \frac{V_{if1} - V_{if2}}{I_s}\frac{I_s}{V_{rf}} = G_{sw}G_{tr} \quad (15)$$

III. SIMULATION VERIFICATION

The harmonic balance simulation is used to verify the analytical expression for mixer's conversion gain. The Angelov model [5] for GaAs HEMT is applied in the simulation. The parameters used in the theoretical calculation based on the analytical expression are also extracted from the Angelov model [5]-[7].

Two types of simulation conditions are setup. One is the fixed LO frequency condition, and the result is shown in Fig. 4. In this case, the full Angelov model is applied to the transistors in the transconductance stage and the simple Angelov model including C_{ds}, C_{gd}, R_d, and R_s, respectively, is applied to the transistors in the switch quad. The calculated conversion gain is almost identical with the simulated ones including R_d and R_s. But it does not fit the curves including C_{ds} or C_{gd} well in high frequency range. The reason is that C_{ds} and C_{gd} in the switch quad are not included in the conversion matrix analysis, and the effect coming from these two parameters depends highly on frequency. Therefore, the theoretical expression for the conversion gain without C_{ds} and C_{gd} in the switch quad is not accurate enough at high frequencies. The fixed IF frequency
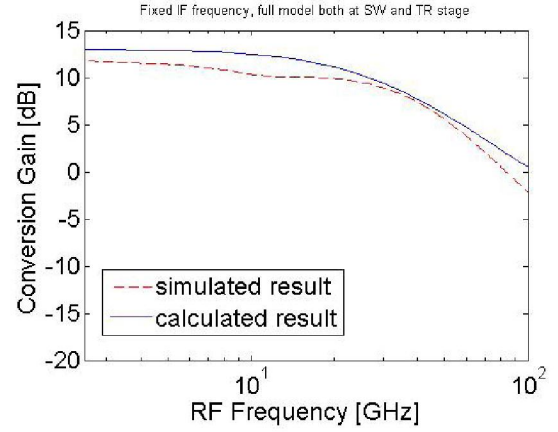

Fig. 5. Comparison of calculated and simulated active Gilbert cell mixer's conversion gain, IF frequency is fixed at 0.5GHz, and C_{ds}, C_{gd}, R_s and R_d of the transistors in switching quad are all included in the simulation.

simulation is also implemented. The result is shown in Fig. 5.

IV. MEASUREMENT VERIFICATION

The theoretical expression of the conversion gain is also verified by chip measurement experiment. A GaAs HEMT active Gilbert cell mixer using OMMIC ED02AH process is applied in the experiments. The measured conversion gains are calibrated simply by taking into account cable losses. The corresponding schematic of the measurement setup is simulated with Agilent ADS. The GaAs HEMT parameters of ED02AH process are also extracted from the ADS ED02AH transistor model for theoretical calculation.

Fixed LO frequency and fixed IF frequency conditions are setup for both chip measurement and ADS simulation. The measurement and simulation results are illustrated in Fig. 6 and Fig. 7 respectively. Although the measured conversion gain varies at high frequencies, the differences among measured, simulated, and theoretical calculated conversion gains are acceptable. The error mainly comes from the inaccuracy of the
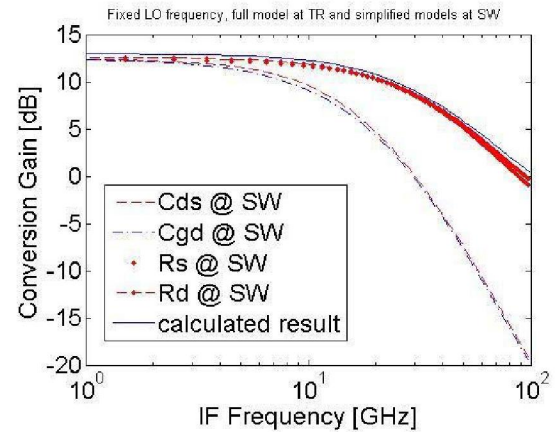

Fig. 4. Comparison of calculated and simulated active Gilbert cell mixer's conversion gain, LO frequency is fixed at 2 GHz, and C_{ds}, C_{gd}, R_s and R_d of the transistors in switching quad are included in the simulation one by one.
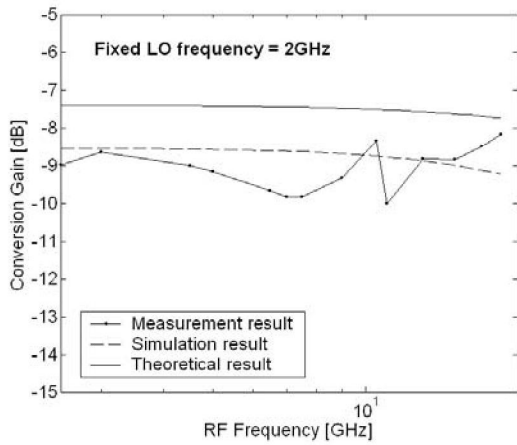
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Fig. 6. Comparison of measured, simulated, and calculated active Gilbert cell mixer's conversion gain, LO frequency is fixed at 2 GHz.
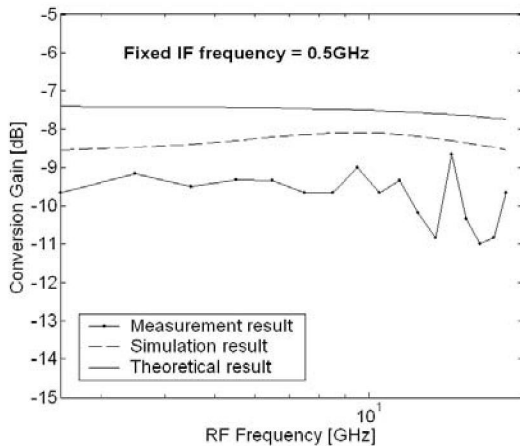

Fig. 7. Comparison of measured, simulated, and calculated active Gilbert cell mixer's conversion gain, IF frequency is fixed at 0.5 GHz.

ADS transistor model and nonlinear small signal circuit model of active Gilbert cell mixers.

V. CONCLUSION

The analytical expression for the GaAs HEMT active Gilbert cell mixer is derived by conversion matrix analysis when the transistor's model is simple enough. The expression is successfully verified by both ADS simulation and chip measurement. The excellent agreement is achieved between calculated and simulated results when the corresponding simple Angelov model is used in the simulation. It is also concluded that the mixer's conversion gain in high frequency domain can be increased by decreasing the effect originating from C_{ds} and C_{gd} of the transistors in the switch quad. Current effort concentrates on including a more complete description of the switch quad.

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