Analysis of HBT Vector Modulator Phase Shifters Based on Gilbert Cell for sub-THz Regimes

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Abstract—This paper concerns with the nonlinearity analysis and compensation of active phase shifters based on vector modulator topologies which utilize Gilbert cell type multipliers. In this respect, both frequency dependence and expansion behavior of such topologies are analyzed. Based on translinear circuits theory, log/anti-log nonlinear equations are transformed to algebraic linear equations and/or multiplication equations, which facilitates solution for such nonlinear equations. To validate the analyses, a 5-bit active phase shifter was designed and simulated in 0.13 μ m HBT technology node. 3D electromagnetic simulation results verify the consistency of the analysis and proposed compensation technique.

Keywords—HBT, active phase shifter, vector modulator, Gilbert cell, translinear analysis, sub-THz, 5G, 6G

I. INTRODUCTION

Beamforming offers several advantages to the new wireless systems; the most important of which is, it enables lower transmit-power per unit block in phased array applications when the link-budget is concerned. It also facilitates multipath links in congested line-of-sight (LOS) link. Spatial scanning capability of such systems empowers the multi-user multi-input-multi-output (MIMO) systems including 5G/6G/IoT [1]. On the other hand, the electronic steerable directional antenna patterns offered by such systems has improved radar and imaging applications as well [2]. Phase shifters (PS) are the key enablers that facilitate such broad applications. There are several different types of PS topologies all of which may be categorized into two major classes namely, “passive” and “active” PS’s. The most important advantage of the passive PS’s is their linearity while they suffer mostly from higher losses and limited phase resolution/variation [2] – [6]. Active PS’s, on the other hand, offer gain as well as much better/wider phase resolution/variation at the cost of nonlinearity of the active devices [1], [7] – [17]. Vector modulators are dominantly utilized to shift the phase in active PS’s. Thus, vector modulators play a critical role in all the abovementioned systems.

As the Gilbert cell structure offers reduced circuit complexity, it is possible to implement it at 6G candidate frequencies, i.e. sub-THz, bands. Moreover, its importance in nearly every in-phase/quadrature (I/Q) mixer, multiplier, variable gain amplifier (VGA), has made it a key building block of the mentioned systems.

In order to model the nonlinearities of active phase shifters based on Gilbert cell, translinear circuit analysis method has been applied to express the log/anti-Log equations governing the circuit operation [18]. In this respect the scheme of vector modulator PS’s along with their operating principals is explained in section II. Section III concerns with the operating principles of Gilbert cells based on translinear circuit analysis

theory. The origin of nonlinear behavior of the mentioned phase shifter circuits is analyzed in section IV. The frequency dependent phase error is also covered in the same section. Simulation results are then presented in section V.

II. VECTOR MODULATOR PHASE SHIFTERS

A. Block Diagram

Fig. 1 illustrates the block diagram of the vector modulator PS’s concept, where the input signal is decomposed into I and Q vector components and applied to the VGAs. Based on the requirements, the I and Q vector signals are weighted by the VGAs and the resultant weighted vectors are then added to construct the phase shifted version of the original input signal.

B. Circuit Level Implementation of the Blocks

The quadrature signal generation (Fig. 1) may be implemented in several various circuit schemes such as poly phase filters [11], hybrid couplers [2], [12], lumped elements [13], [14]. All the mentioned implementations share the same property of being passive in nature, which are hardly ever sources of nonlinearity; unlike the VGAs case.

There are also different circuit topologies for the VGA stages, e.g. programmable differential amplifiers [1], [17], Gilbert cell [15], etc.; all of which utilize differential amplifiers circuitry as the core analog signal processor of the vector modulator PS’s. The presence of transistors, certainly, contribute to the nonlinearity, hence improper constellation point relocation in either compression or expansion manner is inevitable. Thus, the rest of this paper is devoted to HBT differential pair and Gilbert cell analysis.

III. I/Q AMPLITUDE WEIGHTING CIRCUITRY

As mentioned earlier, comprised of $T_1 - T_6$, Gilbert cell has been exploited for the VGAs (Fig. 2). T_7 and T_8 have been utilized to linearize the $\tanh$ property of the i - v characteristic of the differential pair [19]. With I and Q branches sharing the same circuit scheme, Fig. 2 summarizes the whole vector modulator. C_b decouples the RF signal from the control circuitry, thus the T_7 and T_8 carry only DC currents.

Fig. 1. Block diagram of the vector modulator phase shifter along with the weighted I/Q component and the resultant shifted version of the input signal.

$$(I_1')^{-1} = e^{\frac{I_m \sin \omega t}{2\omega C_\pi V_T}} \left(\frac{1}{2C_\pi V_T(1-X)} \underbrace{\int e^{\frac{I_m \sin \omega t}{2\omega C_\pi V_T}} dt}_{ii} + C_o \right) \quad (9)$$

C_o in (9) is a constant which can be defined with respect to initial conditions of the vector modulator circuit. To calculate (9), the terms i and ii , can be expanded in a Fourier series, in which both terms are expressed by the definition of modified Bessel function of order n and argument $(I_m/2\omega C_\pi V_T)$. Since the operating frequency is quite close to f_i , only the fundamental harmony may be of interest, that is the order of the Bessel function can be limited to $n = 1$. Thus,

$$I_1' \propto (1-X) \underbrace{(J_0(x) + 2\omega C_\pi V_T J_1(x))}_{\text{expansive term}} \sin \omega t. \quad (10)$$

Where J_0 and J_1 are the Bessel functions of the zeroth and first order respectively and x is the argument of the Bessel function defined as $(x = I_m/2\omega C_\pi V_T)$.

Equation (7) must be modified in accordance with (10); i.e. all the terms need to be replaced with their corresponding term in (10). Equation (10) approximates to the sub-THz behavioral model of vector modulator PS's that predicts the impact of the parasitic capacitance C_π at different frequencies. Due to the presence of modified Bessel function, the nonlinear mapping of the constellation points in an expansive manner is expected. It is also clear that as the operating frequency approaches zero (10) approaches (5) and (6) and hence (7).

Part of such nonlinearities can be relieved by inductive compensation in the TL loop in proximity of the base-emitter junction capacitances. Since there exist two coupled loops, the inductance element, i.e. L_{cp} , can be shared between the loops shown in Fig. 2. The approximate value of L_{cp} can be determined by the help of Fig. 5 as

$$L_{cp} = \frac{1}{\omega^2 C_\pi}. \quad (10)$$

In order to dimension L_{cp} , the parasitic capacitances present in each primary TL loop of Fig. 2 were placed in Fig. 5. This is starting from collector of T_5 , i.e. T_{5C} , to the base of T_1 , i.e. T_{1B} , where parasitic capacitance of the control circuitry, i.e. C_c , is added to the node, and then to T_{7B} and T_{4B} and back to the T_{5C} . By the same token, other parasitic capacitances may be added to form Fig.5 and from that the equivalent parasitic capacitance can be approximately estimated.

V. SIMULATION RESULTS

In order to verify the proposed compensation method and validation of the extracted equations, a phase shifter was designed which utilizes Gilbert cell, as the core signal processing circuitry, and a 5-bit current DAC, for the sake of level control and constellation point mapping. The circuit was designed in HBT 0.13μm technology node with f_t/f_{max} of 350/550 GHz. In order to track the impact of frequency variations on constellation points, the operating frequency was swept from frequencies much below f_i up to those quite close to f_i . The feasibility of the proposed compensation method is examined at 300 GHz. To account for the impact of the metallic parts, a 3D EM simulation was performed. A simplified Gilbert cell structure is shown in Fig. 6.

Fig. 7 illustrates the simulation result of the designed circuit at 10GHz. It can be seen at low frequencies the response of the phase shift to the control coefficients is totally linear with neither sign of expansion nor constellation points curvature mapping. This is due to that fact that parasitic capacitances have a minor impact at such frequency bands.

Increasing the operating frequency up to 100GHz will demonstrate the first symptoms of the constellation mapping deterioration. This is shown in Fig. 8. It is immediately seen that the constellation points not only have rotated but also expanded over the control range. This is because the parasitic reactive components start to manifest, i.e. ωC_π values are comparable to g_m .

With increasing operating frequency and approaching the f_i of the transistors, the situation tends to get worse. This is quite compatible with what was foreseen by (10). Drastic expansion nonlinear mapping is clearly observed in the constellation point (see Fig. 9).

After applying the proposed compensation method, the nonlinear behavior gets relieved to some extent (see Fig. 10). However, a closer investigation of the constellation points illustrates existence of some nonlinearity (Fig. 11). This is because the proposed compensation technique, compensates mostly the linear average part of the impact of the parasitic capacitances even though they are nonlinear in their nature.

The state trajectory of the transconductance of the transistors at different frequency bands is also very informative. As can be seen from Fig. 12, at frequencies well below f_i , e.g. 10GHz, the transconductance state trajectories hardly show any nonlinear behavior. Nonlinearities as well as drastic phase rotation increase with operating frequency. This is shown in Fig. 13. By the virtue of the proposed compensation method, part of the phase rotation as well as nonlinearities of the state trajectories are averaged out.

Fig. 5. Parasitic capacitances and their equivalence to be compensated by L_{cp} .

Fig. 6. A simplified 3D view of the Gilbert cell utilized in the design.

Fig. 7. Constellation points at low frequencies.

Fig. 8. The signs of constellation deterioration as the frequency increases.

Fig. 9. Total constellation point deterioration at frequencies close to f_f of the transistors, before compensation. The operating frequency is 300 GHz.

Fig. 10. The impact of the proposed compensation technique on the constellation points at 300 GHz.

VI. CONCLUSION

Active phase shifters based on Gilbert cell were analyzed in this paper. Nonlinearity mechanism of such circuits was determined and approximately modeled. Based on the analysis the origin of the nonlinearity was recognized. A compensation

method was proposed and validated based on simulations which proved the concept. To support the simulations, the transconductance state trajectories were presented to illustrate phase rotation and nonlinear distribution of the constellation points.

Fig. 11. Magnified constellation points of the compensated PS.

Fig. 12. State trajectory of the IV of the Gilbert cell at 10 GHz.

Fig. 13. Phase rotation and nonlinearity of the state trajectory of the IV of the Gilbert cell at 300 GHz, before compensation.

Fig. 14. Phase rotation and nonlinearity of the state trajectory of the IV of the Gilbert cell at 300 GHz, after compensation.

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