High-Efficiency InP-based DHBT Active Frequency Multipliers for Wireless Communications

Héctor J. De Los Santos, Daniel D. Nardi, Kevin L. Hargrove, Madjid Hafizi, and William E. Stanchina

Abstract— **We present, for the first time, the performance of AlInAs/GaInAs/InP double heterostructure bipolar transistors (DHBT's) as active frequency multipliers at frequencies of relevance to Wireless Communications applications. In particular, we present results comparing the performance of $X6$ (127 → 762 MHz) and $X4$ (762.5 → 3050.0 MHz) InP- and Si-based (NEC2107) multiplier circuits. A well-known multiplier circuit topology has been chosen as a vehicle, so that we can focus on *active device* comparison. The $X6$ InP-based multiplier exhibits output power and efficiency of +6 dBm and 11%, respectively, compared to +7.2 dBm and 4.8% of the Si-based circuit. The $X4$ InP-based multiplier exhibits output power and efficiency of +3.74 dBm and 8%, respectively, compared to −6.1 dBm and 0.4% of the Si-based circuit. The superior electronic properties of InP-DHBT's enable high-conversion gain/highly DC-efficient multipliers, however, their nonexponential I_C-V_{BE} characteristic limits the maximum obtainable conversion gain at high-order frequency multiplication.**

I. Introduction

A new era of wireless communications systems has emerged [1]. Such systems are built around the concept of a handset which is: highly portable, low power, and low cost [2]. Systems solutions in which the local oscillator (LO) function is implemented off-chip [2]–[5], demand a rather efficient implementation of the LO chain. The key to a power-efficient LO chain lies in the frequency multiplier. The widespread use of transistor frequency multipliers in microwave engineering has its roots in, at least, two important facts. First, their ability to provide amplification may result in a multiplication efficiency exceeding the fundamental limit of $1/n^2$, where n is the order of multiplication [6], [7]. Secondly, they require a relatively low input signal level [6]. One key parameter that limits the performance of transistor frequency multipliers to relatively low input frequencies and multiplication orders, n, is the transistor's cutoff frequency, f_T. As the current conversion efficiency for output harmonic n is approximately f_T/nf_0 [8], it is clear that one way to enhance the efficiency is by using a high-f_T device. While AlInAs/GaInAs/InP HBT's, which have been demonstrated to exhibit typical f_{MAX}'s ∼ 90 GHz and f_T's ∼ 65 GHz [11], were originally develop to target millimeter-wave applications, new emphasis on *dual-use* makes it a prime candidate for wireless communications applications, in particular, frequency multipliers, in light of its excellent high-frequency properties. In this letter we present results pertaining the exploratory development of AlInAs/GaInAs/InP DHBT-based active frequency multipliers, and compare their performance to conventional silicon-based circuits.

II. Approach

The formal design of microwave frequency multipliers encounters a number of difficulties, not the least of which, is the need for deriving a multi-frequency, multi-power, multi-bias large-signal device model [6]. Such an approach is beyond the scope of this letter, as our emphasis is on investigating the performance of *developmental* AlInAs/GaInAs/InP DHBT's and thus *no effort* has been directed toward developing optimized circuits for these devices, rather we have focused on the prospect of *replacing* Si-based multipliers *currently in production*, which utilize a circuit like that shown in Fig. 1 [9].

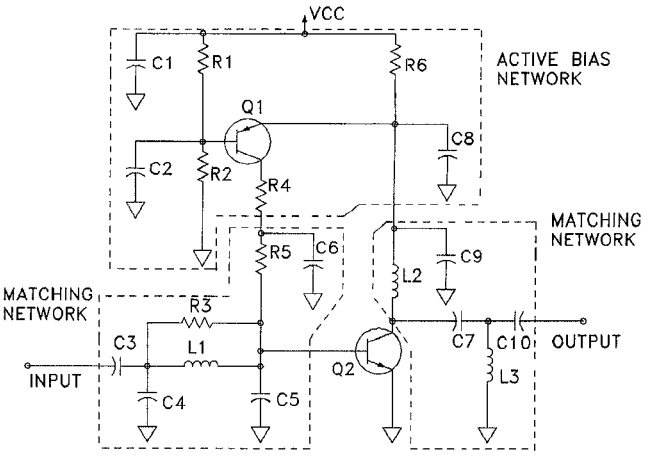

Fig. 1. Frequency multiplier circuit schematic. The circuit was implemented in standard MIC technology.

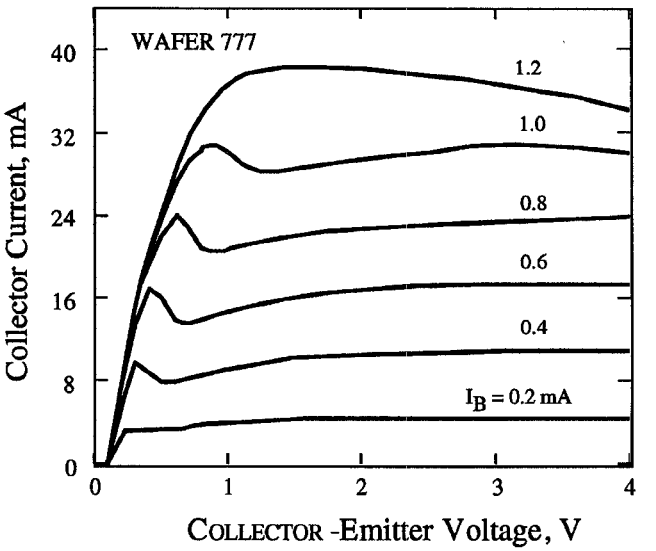

Fig. 2. Typical InP-DHBT common-emitter characteristics. Emitter area = 2 × 20 μm^2.

The DHBT's were die-attached using Silver epoxy to a 70 mil ceramic package. The device contacts were wirebonded to the package leads using an approximately 40 mil long/1 mil diameter Gold wire. A typical current-voltage characteristic of the DHBT device in the common-emitter configuration is shown in Fig. 2. At low VCE, as can be seen, the device exhibits what appear to be negative resistance regions. This has been found to result from an incompletely-graded base-collector conduction band discontinuity [10]. Due to the discontinuity, the collector current is not an exponential function of V_{BE}, as is usual in diffusion-controlled base-collector transport, but rather, varies as $I_C \sim V_{BE}^2 \exp(-b/V_{BE})$, as in tunneling-controlled transport [10]. Because of this, most of the harmonic power is expected to be concentrated in the lower harmonics. The DHBT devices used exhibited a typical $f_{MAX} \sim 90$ GHz, $f_T \sim 65$ GHz,

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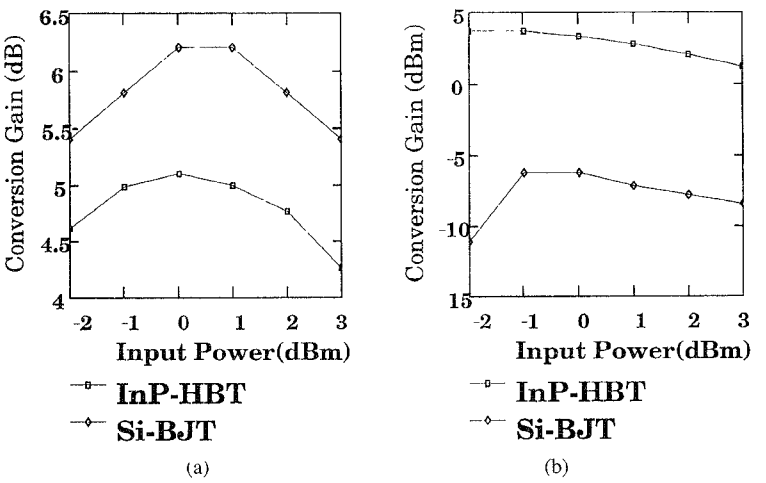

Fig. 3. Measured conversion gain versus input power (a) $X6$ multipliers (b) $X4$ multipliers.

TABLE I
NOMINAL PERFORMANCE OF TRANSISTOR FREQUENCY MULTIPLIERS

Nominal Pin=+1dBm Nominal Temp.=+23°C	*X6* (fin=127 MHz/ fout=762 MHz)	*X4* (fin=762.5 MHz/ fout=3050 MHz)
Device	Si-BJT / InP-DHBT	Si-BJT / InP-DHBT
Nominal Output Power	+7.2dBm / +6.0 dBm	-6.1dBm/ +3.74dBm
0°C - 60°C Power Variation	0.2dB / 0.34dB	0.6dB / 0.66dB
Input Return Loss	20.6dB / 16.6dB	14.4dB / 16.2dB
Output Return Loss	12.1dB / 2.45dB	8.3dB / 4.2dB
DC Power Consumption	110mW / 36mW	62.3mW / 29mW

collector-base (with emitter open) breakdown voltage, $BV_{CBO} \sim 30$ V, and current gain, $\beta \sim 30$ [11]. The silicon bipolar transistors utilized were commercially-available packaged NEC2107 with typical $f_T \sim$ 5-6 GHz, and $\beta \sim 100$.

To carry on the evaluation $X6$ and $X4$ multiplier circuits were built, Fig. 1. First, with the control active device Q2 being a Si NEC2107, the circuits were submitted to our production line where they were iteratively tuned and tested for best input/output return loss, output power, DC power and temperature performance, as is routinely done in production. Next, the active Si devices were *replaced* by AlInAs/GaInAs/InP DHBT's, and the same tuning/testing procedure was followed. The results for the Si-based circuits, which are consistent with the best customarily obtained in our production facility, together with those of the DHBT-based circuits are shown in Fig. 3 and Table I, and discussed next.

III. RESULTS AND DISCUSSION

The measured conversion gain versus input power, and nominal (Pin = +1 dBm) performance of $X6$ (fin = 127 MHz and fout = 762 MHz), and $X4$ (fin = 762.5 MHz and fout = 3050 MHz) multipliers is shown in Fig. 3, and Table I, respectively. The conversion gain of the Si-BJT-based $X6$ multiplier is seen to be consistently greater, by about 1 dB, than that of the InP-DHBT-based, although it exhibits a lower nominal DC efficiency, i.e., 4.8% versus 11%. Since the frequency is low-enough, compared to the f_T's, the major factor limiting the conversion efficiency is the I_C-V_{BE} transfer characteristic, thus the lower relative conversion gain of the InP-DHBT-based multiplier is attributed to its less nonlinear, *nonexponential*, transfer curve [10], which is ill-suited for achieving high-order multiplication. At higher frequencies and lower order multiplication, e.g., the $X4$ case, it is seen in Fig. 3, that the InP-DHBT-based multiplier is the one exhibiting the higher conversion gain, together with a higher DC efficiency, i.e., 0.4% versus 8%. Here, the InP-DHBT-based circuit owes its superior performance to both the lower-order multiplication, which is favored by its transfer curve, and its vastly greater f_T capability. The higher DC efficiency exhibited in both cases by the InP-DHBT based multipliers is due to the InP-DHBT's larger transconductance and lower threshold V_{BE} [11] compared with the Si-BJT device.

IV. CONCLUSION

We have presented, for the first time, the performance of developmental AlInAs/GaInAs/InP DHBT-based active frequency multipliers, and have compared it to that of conventional silicon-based circuits, for potential application in low-power wireless communications. While the superior electronic properties of InP-DHBT's, in particular, their higher transconductance, lower threshold V_{BE}, and higher f_T give rise to high-conversion gain/highly DC-efficient frequency multipliers, the nonexponential nature of their transfer I_C-V_{BE} poses a limitation on the maximum obtainable conversion gain at high-order frequency multiplications.

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Analysis of Slot-Coupled Double-Sided Cylindrical Microstrip Lines

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***Abstract*— The problem of double-sided cylindrical microstrip lines coupled through a rectangular slot in the common cylindrical ground plane is studied using the reciprocity theorem and a moment-method calculation. The oretical results of the S parameters for the slot-coupled cylindrical microstrip lines are calculated and analyzed. Experiments are also conducted to verify the theoretical results.**

I. Introduction

Slot-coupled double-sided microstrip lines have recently found applications in the design of directional couplers [1], which are useful for beam-forming networks, multiport amplifiers, and other important microwave and millimeter-wave circuits. Several related studies have also been reported [2]–[4], in which the slot-coupled microstrip lines on a planar geometry are treated. In this paper we present an analysis of the slot-coupled microstrip lines in a cylindrical structure; i.e., the slot-coupled double-sided cylindrical microstrip lines. This new structure is useful for the design of conformal printed circuits on cylindrical surfaces. To perform the analysis we apply the reciprocity theorem and use the exact Green's functions for the grounded cylindrical substrate in a moment-method calculation for the unknown slot electric fields. The formulation of the S parameters for the slot-coupled cylindrical microstrip lines is presented, and theoretical results are calculated and discussed.

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II. Theoretical Formulation

Fig. 1 shows the geometry under consideration. The inside cylindrical microstrip line [5], [6] is treated as a feed line, and the outside cylindrical microstrip line [7] is the coupled line. These two lines are assumed to be infinitely long and are coupled through a rectangular slot of dimensions $L \times W$ in the common cylindrical ground plane of radius b. The widths of the feed and coupled lines are $W_f(=2a\phi_f)$ and $W_c(=2c\phi_c)$, respectively. The feed substrate has a thickness h_f and a relative permittivity ε_f; the coupled substrate has a thickness h_c and a relative permittivity ε_c. The inner ($\rho < a$) and outer ($\rho > c$) regions are air with permittivity ε_0 and permeability μ_0. To begin with, we assume that the input power at port 1 is 1 watt and the microstrip lines are propagating quasi-TEM waves [2], [8]. And, by considering that the coupling slot is narrow ($L \gg W$), the electric field in the slot can be approximated as

$$\vec{E}^s = \hat{z} \sum_{q=1}^{N} V_q e_q^s(\phi, z) = \hat{z} \sum_{q=1}^{N} V_q \frac{1}{W} f_p^s(\phi - \phi_q), \quad |z| < \frac{W}{2}, \quad |\phi| < \phi_s \tag{1}$$

with

$$f_p^s(\phi) = \begin{cases} \dfrac{\sin k_e(b\phi_h - b|\phi|)}{\sin k_e b\phi_h}, & |\phi| < \phi_h, \\ 0, & |\phi| > \phi_h \end{cases}$$

$$\phi_q = \left[\frac{2q}{N+1} - 1\right]\phi_s,$$

$$\phi_h = \frac{2\phi_s}{N+1},$$

$$\phi_s = \frac{L}{2b},$$

$$k_e = k_0\sqrt{\frac{\varepsilon_f + \varepsilon_c}{2}},$$

$$k_0 = \omega\sqrt{\mu_0\varepsilon_0}$$

where $f_p^s(\phi)$ is a piecewise sinusoidal (PWS) basis function, ϕ_q is the center point of the qth expansion function, $b\phi_h$ is the half-length of the PWS function, and V_q is the unknown coefficient of the qth expansion function.

To solve for V_q, two boundary conditions are applied; one is the continuity of the tangential magnetic field at the slot position, and the other is that the tangential electric field must be zero on the coupled line. By considering the boundary condition that the tangential magnetic field must be continuous in the slot, we have

$$H_\phi^f + H_\phi^{sf} = H_\phi^c + H_\phi^{sc} \tag{2}$$

where H_ϕ^f and H_ϕ^c are, respectively, the fields contributed from the feed line and the coupled line in the absence of the slot; H_ϕ^{sf} and H_ϕ^{sc} are, respectively, the fields at $\rho = b^-$ and $\rho = b^+$ from the slot. By deriving the appropriate Green's functions for the cylindrical structure studied here, the magnetic fields in (2) can be expressed as

$$H_\phi^f = (1 - \Gamma)h_\phi^f \tag{3}$$

$$H_\phi^{sf} = \iint_{S_a} G_{\phi\phi}^{HMf}(b^-, \phi, z)M_\phi^{sf}\, ds_0 \tag{4}$$

$$H_\phi^c = \int_{-\infty}^{\infty}\int_{-W_c/2}^{W_c/2} G_{\phi z}^{HJc}(b^+, \phi, z)J_z^c\, ds_0 \tag{5}$$

$$H_\phi^{sc} = \iint_{S_a} G_{\phi\phi}^{HMc}(b^+, \phi, z)M_\phi^{sc}\, ds_0 \tag{6}$$