A Novel Heterojunction Bipolar Transistor VCO Using an Active Tunable Inductance

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***Abstract*—This paper reports on the results of a novel HBT Voltage Controlled Oscillator (VCO) that incorporates a bias-tunable active inductor. This development benchmarks the first demonstration of a tunable active inductance controlled HBT VCO of this type. The active inductor topology can obtain inductances of up to 10 nH at frequencies up to 10 GHz and can be bias-tuned over a similar range. VCO's were designed at 4 and 10 GHz by implementing different active inductance values. A 29% frequency tuning range from 3.54 to 4.73 GHz was obtained for the 4-GHz design. The output power varied from −1.2 to −0.2 dBm, respectively—less than 1 dB variation over the tuning range. The phase noise is −70 dBc/Hz at 100 kHz offset from the carrier. The new VCO topology can be implemented without the use of backside vias or microstrip-matching components and can be realized in a compact 1.0 × 0.7 mm^2 area. The variable active inductance VCO topology provides a compact, high performance alternative to the analog multi-vibrator oscillator circuit that generally has a smaller tuning capability and worse phase noise performance at microwave frequencies.**

I. Introduction

Active-inductor topologies have previously been developed in order to economically implement monolithic inductive elements at microwave frequencies [1], [2]. These developments have led to amplifier circuits that incorporate frequency-dependent active loads. The active loads provide gain peaking in amplifier circuits, which enhances the gain-bandwidth performance. Other uses of active inductors have been in HBT active feedback amplifiers [3]. In these amplifiers, the active inductance provides proper phasing in order to obtain positive feedback. The present work describes the use of an active inductor to provide a tunable resonator tank circuit that is integrated with an HBT oscillator. The bias-tuning properties of the active inductor provides a means of controlling the oscillation frequency. The resulting circuit is a very compact analog implementation of a microwave bipolar VCO that can be used as a miniature, low cost LO source for receivers or in light-wave Phased Lock Loop (PLL) applications.

II. Tunable Active Inductor VCO Design

In conventional microwave oscillators, an inductive element is normally used to resonate with the intrinsic capacitances of a transistor to set the frequency of oscillation. In addition, a means of tuning the resonant frequency is needed to realize a voltage control oscillator. Voltage frequency control is usually implemented with a varactor diode (variable capacitance) or by using the bias-dependent properties of the device intrinsic capacitances. However, these implementations result in either limited frequency tuning range or varying output power with frequency tuning. The active inductor implementation described in this work can provide both wide band frequency tuning as well as output power insensitivity.

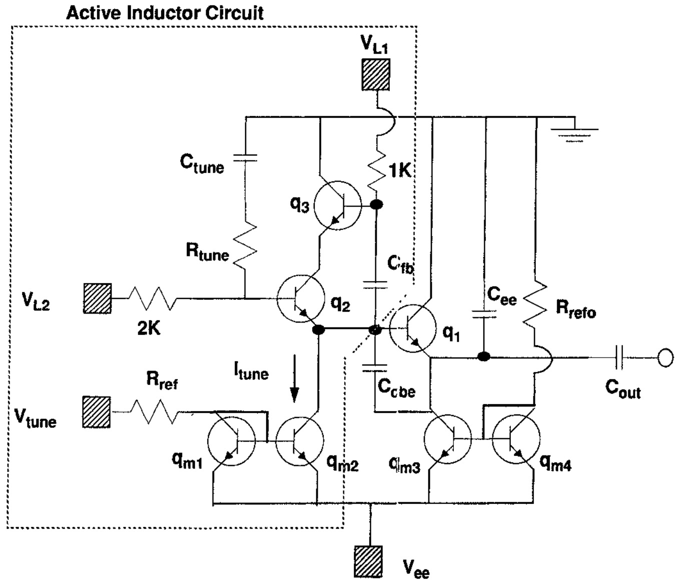

Fig. 1. Detailed schematic of the HBT active tunable inductor VCO.

The full HBT VCO design schematic is shown in Fig. 1. Transistor Q_1 is the main HBT transistor of the oscillator, which provides a negative resistance at its base terminal. The value of capacitor $C_{\rm ee}$ is adjusted to induce negative resistance at the base of Q_1 for the frequency band of interest. Capacitor $C_{\rm out}$ lightly couples the oscillation energy to the desired output load. Capacitance $C_{\rm cbe}$ is an external capacitance that supresses the oscillation frequency sensitivity (phase noise) to bias changes in transistor Q_1. Transistors Q_{m3}, Q_{m4} and resistor $R_{\rm refo}$ provide a convenient means of biasing Q_1. The rest of the circuit is comprised of an active inductor network that is similar to a previously developed lossless active inductor in MESFET technology [1]. Transistors Q_2 and Q_3 comprise the active portion of the active inductor. An inductive impedance looking into the emitter of Q_2 is

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synthesized and is related to the magnitude of R_{tune} and the bias current I_{tune}. Capacitor C_{fb} helps extend the inductive impedance to higher frequencies, while C_{tune} provides a dc block. A current-mirror is used to set up a voltage-to-current converter for controlling the bias current I_{tune} of the active inductor network.

The impedance looking into the emitter of Q_2 is dependent on both R_{tune} and I_{tune}. A relationship expressing the active impedance of this inductor topology in terms of MESFET device model parameters has previously been derived [1]. For this HBT implementation, a similar derivation is provided below which expresses the active impedance in terms of HBT device parameters. Assuming the simplest hybrid-π model where r_e, r_b, r_c, and $C\mu$ are set to zero, the active impedance is expressed as

$$Z_{\text{ind}}(\omega) = \frac{1 + j\omega C_{\pi 2} R_{\text{tune}}}{G_{m2} + j\omega \left[C_{\pi 2} - C_{\pi 3}\left(\frac{G_{m2}}{G_{m3}}\right) + \omega^2 C_{\pi 3}\left(\frac{C_{\pi 2} C_{\pi 3}}{G_{m3}^2}\right)\right]} \quad (1)$$

where G_{m2} and $G_{m3}, C_{\pi 2}$, and $C_{\pi 3}$ are the transconductances and input capacitances of HBT transistors Q_2 and Q_3, respectively. R_{tune} is a design resistance used to set the nominal value of the active inductor.

If Q_2 and Q_3 are the same size, then the expression can be further reduced:

$$Z_{\text{ind}}(\omega) \approx \frac{1 + j\omega C_\pi R_{\text{tune}}}{G_m + j\omega C_\pi \left(\frac{\omega C_\pi}{G_m}\right)^2} \quad (2)$$

if $(f/f_T)^3 \ll 1$ then,

$$Z_{\text{ind}}(\omega) \approx \frac{1}{G_m} + \frac{j\omega C_\pi R_{\text{tune}}}{G_m} = R + j\omega L \quad (3)$$

where R and L can be further broken down to include bias current dependence:

$$R = \frac{1}{G_m} = \frac{n \cdot kT}{I_c \cdot q} \quad (4)$$

$$L = \frac{C_\pi(I_c) \cdot R_{\text{tune}}}{I_c \cdot q} \cdot nkT \quad (5)$$

By inspection, we can see that L is directly proportional to R_{tune}, however it is not obvious how L is dependent on I_c. The dependence of C_π on I_c is given by $C_\pi = C_{\text{be}} + I_c \cdot G_m \cdot n/kT$. This additional information allows us to conclude that the inductance, L, will increase with a decreasing collector current.

Small-signal simulations indicate that $2 \times 10\ \mu\text{m}^2$ HBT's can generate an inductance of as much as 10 nH, and can be bias-tuned over a similar range of inductances. The active inductor Q-factors can be as much as 10–20 for the HBT active inductors. Libra simulations of the VCO circuit suggest that this novel VCO design can cover a 100 MHz to greater-than-10-GHz frequency range, with a tuning capability of as much as 35–40%. This is determined by analyzing the small-signal negative resistance and reactive slope of the VCO and resonator.

Fig. 2 shows a microphotograph of a fabricated 4-GHz VCO MMIC. The MMIC was fabricated using InAlAs/InGaAs HBT technology that has been previously described [4], [5].

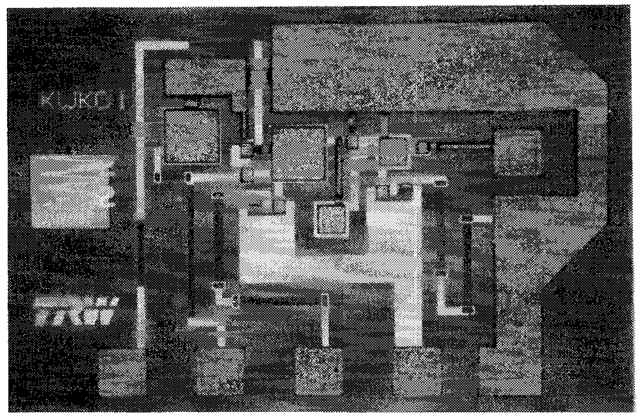

Fig. 2. Micro-photograph of the fabricated chip. The chip size is 1.0×0.7 mm^2.

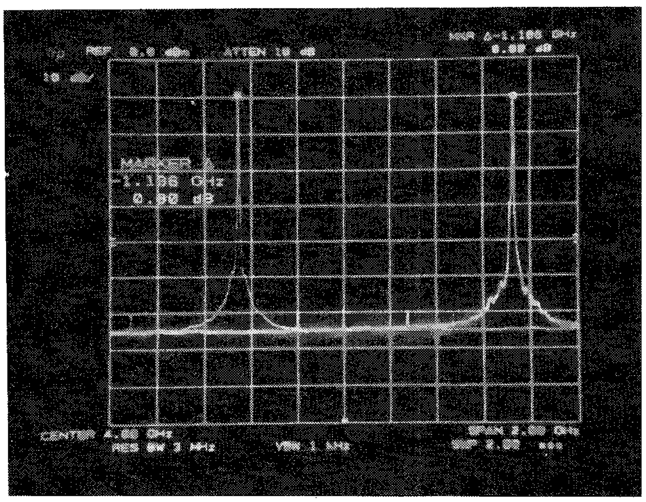

Fig. 3. Spectral output of the 4-GHz VCO: i) -1.2-dBm output power at 3.54 GHz, and ii) -2.0-dBm output power at 4.73 GHz.

The design incorporates HBTs, resistors, and capacitors in an analog layout arrangement. No backside vias or microstrip components are required. This allows the circuit to be easily integrated in a phase-lock-loop IC. The total chip size of the proto-type design is 1.0×0.7 mm^2, however much of this area is consumed by text, vacant area, and unnecessary ground and rf probe pads. The integrated VCO can be optimized to consume an area of less than 0.5×0.4 mm^2.

III. Measured VCO Results

Fig. 3 shows the measured spectral output of a 4-GHz VCO. This spectrum illustrates the measured tuning range of the VCO, as well as the output power insensitivity. The VCO was tuned from 3.54 to 4.73 GHz. This is a tuning range of 29%. The output power varied from -1.2 dBm to -2 dBm, respectively, which is less than 1-dB variation. The phase noise was directly measured on an HP8566 spectrum analyzer. For the 4-GHz VCO, a SSB phase noise of -70 dBc/Hz in a 1-Hz band was measured at 100-kHz offset from the 4-GHz carrier.

A 10-GHz VCO using this novel topology was also designed and measured. The only difference between the 4- and 10-GHz designs was the value of the active inductance. For the

TABLE I
SUMMARY OF 4- AND 10-GHz VCO MESUREMENTS

Parameter	Fo = 4.73 GHz (4.0 GHz Design)	Fo = 3.54 GHz (4.0 GHz Design)	Fo = 10.6 GHz (10 GHz Design)
Iee(Vee=-6V)=	58 mA	58 mA	58 mA
Vtune (V) =	0 V	-3.5 V	0 V
Pout (dBm) =	-2.0 dBm	-1.2 dBm	-2.5 dBm
Phase Noise =	-70 dBc/Hz*	-70 dBc/Hz*	–60 dBc/Hz*

* approximate manual measurement on Spectrum Analyzer

10-GHz VCO design, a lower R_{tune} resistance was used to obtain a lower active inductance. This shifts the VCO resonant frequency up from 4 to 10 GHz. At this frequency a similar frequency tuning range to the 4-GHz VCO is expected. Table I gives a summary of the measured performance of the two VCO designs. This table shows that the X-band VCO achieves an output power of -2.5 dBm at a frequency of 10.6 GHz and a corresponding phase noise of ≈ -60 dBc/Hz at 100-kHz offset.

IV. CONCLUSION

A novel HBT VCO design has been developed that uses active bias-tunable inductors. A 4-GHz VCO demonstrated a frequency tuning range of 29% with an output power ranging from -1.2 to -2 dBm. The measured phase noise was -70 dBc/Hz at 100-kHz offset based on a manual measurement. A 10.6-GHz VCO was also demonstrated with -2.5 dBm of output power and a phase noise of ≈ -60 dBc/Hz at 100-kHz offset. Furthermore, this topology can be fabricated without backside vias or microstrip components, making this design very compact and low cost. This miniature, low-cost analog design can be used at microwave frequencies in PLL applications where analog multi-vibrator VCO topologies begin to reach their performance limits.

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