two-tone input at 1.79 and 1.81 GHz. The conventional FFAMP provides a 16-dB cancellation on the third-order intermodulation (IM3) distortions compared to the main amplifier without linearization. A similar degree of cancellation is observed for other in-band distortion products. When the phase equalizers were used, a further improvement of 4 dB on IM3 cancellation is achieved. Fig. 8 shows the IM3 cancellation over the frequency band of interest obtained from several two-tone measurements. The proposed FFAMP attains better characteristic over the whole bandwidth with a maximum of 6-dB cancellation improvement compared to the conventional design.

V. Conclusions

A new topology of FFAMP with improved wide-band performance has been presented in this paper. Phase equalizers were designed to approximate the nonlinear dependence of phase on frequency of the main and error amplifiers, reducing the nonlinear phase imbalances within the cancellation loops. A 1.7–1.9-GHz FFAMP incorporating the phase equalizers was designed, fabricated, and tested. The proposed FFAMP achieved an improved cancellation characteristic over the whole bandwidth, with a maximum of 6-dB IM3 distortion improvement, compared to the conventional design. This demonstrates the wide-band distortion characteristic of FFAMPs can be improved with the use of phase equalizers.

References

[1] J. K. Cavers, "Adaptive behavior of a feedforward amplifier linearizer," *IEEE Trans. Veh. Thechnol.*, vol. 44, pp. 31–39, Feb. 1995.

[2] S. Narahashi and T. Nojima, "Extremely low-distortion multi-carrier amplifier—Self-adjusting feedforward amplifier," in *Proc. IEEE Int. Commun. Conf.*, June 1991, pp. 1485–1490.

[3] Y. K. G. Hau, V. Postoyalko, and J. R. Richardson, "Sensitivity of distortion cancellation in feedforward amplifiers to loops imbalances," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Denver, CO, June 1997, pp. 1695–1698.

[4] J. P. Dixon, "A solid-state amplifier with feedforward correction for linear single-sideband applications," in *Proc. IEEE Int. Commun. Conf.*, 1986, pp. 728–732.

[5] PST Inc., "High power feed forward amplification systems," *Microwave J.*, vol. 37, no. 2, pp. 128–133, Feb. 1994.

[6] Y. K. G. Hau, V. Postoyalko, and J. R. Richardson, "Compensation of amplifier nonlinear phase response to improve wideband distortion cancellation of feedforward amplifiers," *Electron. Lett.*, vol. 33, no. 6, pp. 500–502, Mar. 1997.

[7] S. O. Scanlan and J. D. Rhodes, "Microwave allpass networks—Part I," *IEEE Trans. Microwave Theory Tech.*, vol. 16, pp. 62–71, Feb. 1968.

[8] K. Konstantinou, P. Gardner, and D. K. Paul, "Optimization method for feedforward linearization of power amplifiers," *Electron. Lett.*, vol. 29, no. 29, pp. 1633–1635, Sept. 1993.

Design of a Low-Supply-Voltage High-Efficiency Class-E Voltage-Controlled MMIC Oscillator at C-Band

Frank Ellinger, Urs Lott, and Werner Bächtold

***Abstract*—In this paper, a monolithically integrated voltage-controlled class-E tuned oscillator for C-band has been designed and measured. Large-signal optimization was performed using analytically calculated starting values to reach high efficiencies at ultra-low supply voltages down to 0.9 V. The range of the tuning voltage is from 0 to the supply voltage. With a supply voltage of 1.8 V, an output power of 6.5 dBm, an efficiency of 43%, and a tuning range of 150 MHz is achieved at a center frequency of 4.4 GHz. With a supply voltage of only 0.9 V, the efficiency is 36%, with an output power of 1.1 dBm, and a tuning range of 80 MHz at a frequency of 3.6 GHz. Chip size is less than 1 mm^2.**

***Index Terms*—MESFETs, MMICs, oscillators.**

I. Introduction

For portable communications equipment, the power consumption has to be minimized. In receivers especially, there is the trend to ever lower supply voltages. The RF section of these receivers consists mainly of low-noise amplifiers (LNAs), mixers, and oscillators. LNAs operating at C-band have been previously reported using supply voltages below 1.5 V [1]. A variety of passive mixers that require no dc supplies have been reported. However, there are few references available on oscillators using low supply voltages. Generally, the efficiency (RF/dc power) of oscillators reduces with decreased supply voltages, thus, most designs require relatively high supply voltages [2]–[4]. Table I shows a summary of oscillators at microwave frequencies with state-of-the-art efficiencies. The lowest supply voltage was 2.0 V, as reported in [5]. An efficiency (η) of 61% at L-band has been reached with a 0.25-μm high electron-mobility transistor (HEMT) process. However, note that except for [3], none of the listed high-efficiency oscillators has frequency tuning.

The purpose of this paper was the design of a highly efficient class-E C-band voltage-controlled oscillator (VCO) with ultra-low supply voltages from 1.8 to 0.9 V, which required minimum chip size.

Large-signal optimization was performed using starting values that were analytically calculated by using simple equivalent circuits. A standard 0.6-μm E/D MESFET GaAs foundry process (Triquint TQTRx, Hillsboro, OR) was used for the design.

II. Approximate Analysis

The major parts of the circuit are the class-E output, the feedback in the source, and the resonator.

A. Class-E Output and Bias

Fig. 1 shows the equivalent class-E output network of the VCO. $R_{\rm eff}$ is the effective load resistance, which is the sum of the load resistance and the resistive parasitics. $V_{\rm eff}$ is the effective dc voltage, which is determined by the saturation voltage, the source voltage drop of the bias network, and the resistive losses of the FET. $L_{\rm eff}$ is the class-E inductance L_x, including the output bondwire $L_{\rm bond} (L_{\rm eff} = L_x + L_{\rm bond})$.

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TABLE I
MICROWAVE OSCILLATORS WITH STATE-OF-THE-ART EFFICIENCIES

Supply voltage	f_{osc}	η	Class	Transistor	Ref.
6.5V	5GHz	59%	E	0.5μm GaAs MESFET	[2]
5.5V	1.6GHz	67%	F	0.7μm GaAs MESFET	[3]
5V	2GHz	20%	A	0.8μm GaAs MESFET	[4]
2V	1.8GHz	61%	F	0.25μm GaAs HEMT	[5]
1.8V	**4.4GHz**	**43%**	**E**	**0.6μm GaAs MESFET**	**This work**
0.9V	**3.6GHz**	**36%**			

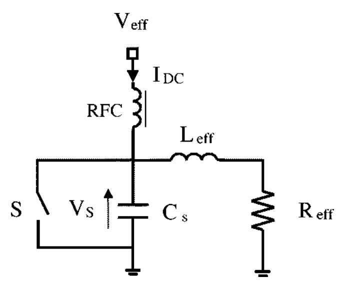

Fig. 1. Equivalent class-E output circuit of the VCO including bias.

The required class-E dc current can be approximately calculated as follows [6]:

$$I_{\rm DC} \approx \frac{2V_{\rm eff}}{g^2 R_{\rm eff}} \tag{1}$$

where g is 1.862 [6]. The class-E capacitance C_S can be estimated with the following relation [6]:

$$C_S \approx \frac{2}{g^2 \omega_{\rm osc} R_{\rm eff} \pi} \tag{2}$$

where $\omega_{\rm osc}$ is the oscillation frequency. The required inductance L_x can be calculated as follows ($L_{\rm eff}$ see [6]):

$$L_x \approx L_{\rm eff} - L_{\rm bond} \approx \frac{1.153 R_{\rm eff}}{\omega_{\rm osc}} - L_{\rm bond}. \tag{3}$$

B. Feedback

To make the active FET unstable, we use capacitive feedback in the source to present a negative resistance at the gate. Fig. 2 shows the equivalent circuit of the FET, including the feedback capacitance C_f. $C_{\rm gs}$ is the gate–source capacitance, and g_m is the transconductance of the FET. The input impedance Z_i at the gate can be calculated using the equivalent circuit of Fig. 3

$$Z_i = -\frac{g_m}{\omega_{\rm osc}^2 C_f C_{\rm gs}} - j\,\frac{C_f + C_{\rm gs}}{C_f C_{\rm gs} \omega_{\rm osc}}. \tag{4}$$

The required feedback capacitance for a certain negative resistance ($\mathrm{Re}\{Zi\}$) at the gate can be calculated using (4)

$$C_f \approx -\frac{g_m}{\omega_{\rm osc}^2 C_{\rm gs} \mathrm{Re}\{Z_i\}}. \tag{5}$$

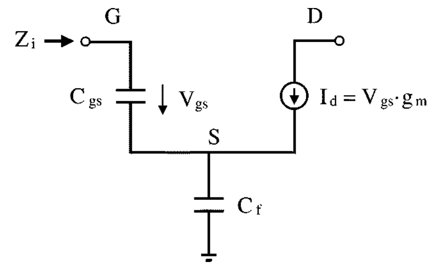

Fig. 2. Equivalent circuit of the FET including the feedback capacitance.

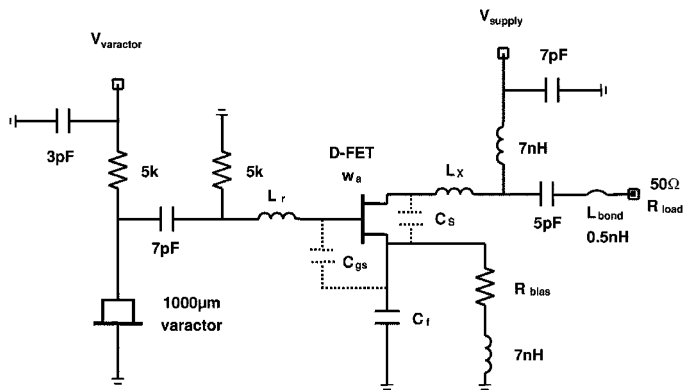

Fig. 3. Circuit topology of the class-E tuned VCO, $C_{\rm gs}$ is the gate–source capacitance, C_s is the output capacitance of the FET.

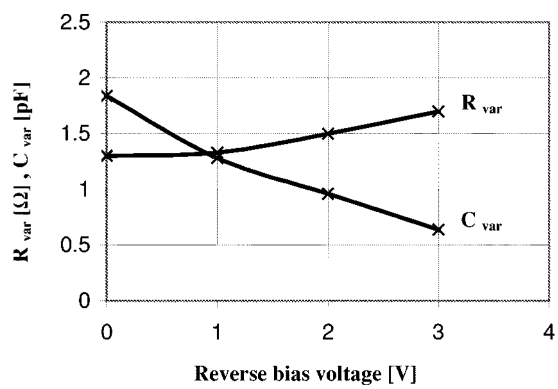

Fig. 4. Diode characteristics of the 1000 μm varactor, extracted from S-parameter measurements.

Small values of C_f increase the negative feedback and ensure oscillation over a larger bandwidth, but decrease the efficiency. An optimum tradeoff in the band of interest has to be found.

C. Resonator

To set the frequency of oscillation, we use an *LC* resonator, consisting of an inductance L_r and a varactor with a capacitance $C_{\rm var}$. C_i is the resulting reactive impedance ($\mathrm{Im}\{Z_i\}$) at the gate of the FET, including the source feedback. The frequency of oscillation can be estimated with the following relation:

$$\omega_{\rm osc} \approx \frac{1}{\sqrt{L_r \left(C_{\rm var}^{-1} + C_i^{-1}\right)^{-1}}} \tag{6}$$

with

$$C_i = \frac{1}{\omega_{\rm osc} \mathrm{Im}\{Z_i\}} = \frac{C_{\rm gs} C_f}{C_{\rm gs} + C_f}. \tag{7}$$

TABLE II
COMPARISON OF IDEALIZED CALCULATIONS AND PSPICE OPTIMIZATION ($V_{\mathrm{supply}} = 1.2$ V, $f_{\mathrm{osc}} = 4$ GHz), FOR CALCULATIONS THE FOLLOWING VALUES WERE ASSUMED: $R_{\mathrm{eff}} \approx 60\,\Omega$, $V_{\mathrm{eff}} \approx 0.8$ V, $g_m \approx 40$ mS, $C_{\mathrm{gs}} \approx 300$ fF and $C_{\mathrm{var}\,(V\mathrm{varactor}=0.6\,\mathrm{V})} \approx 1.5$ pF

Elements	Idealized calculations	Pspice Optimization
I_{DC}	7.7 mA [eq. (1.1)]	5 mA
R_{bias}	45 Ω (from I_{DC})	55 Ω
C_s	122 fF [eq. (1.2)]	150 fF
w_a	244 μm (from C_S)	300 μm
L_{eff}	2.8 nH [eq. (1.3)]	3.5 nH
C_f	1.1 pF [eq. (2.2)]	1.3 pF
L_r	7.6 nH [eq. (3.1)]	5.5 nH

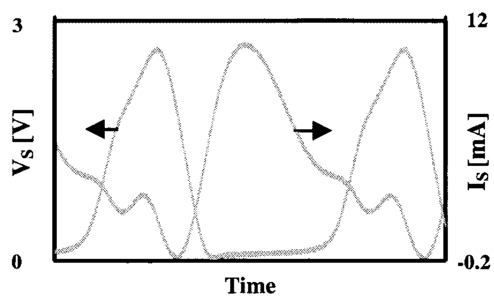

Fig. 5. Simulated waveforms (V_S, I_S) of the Class-E tuned VCO, $V_{\mathrm{supply}} = 1.2$ V, $V_{\mathrm{varactor}} = 0.6$ V.

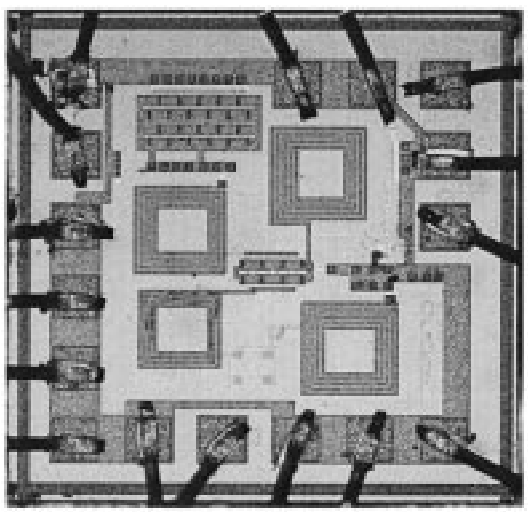

Fig. 6. Microphotograph of the MMIC chip (0.9 × 0.9mm^2) bonded in a test package.

III. CIRCUIT DESIGN

The circuit topology of the VCO is shown in Fig. 3. To minimize chip size, a relatively small dc-feed impedance (dc-feed inductance of 7 nH) and no additional filtering of the harmonics at the output is used. The FET is operated in a self-biased single supply mode. To generate a negative gate–source voltage for the required class-E bias point, the source was dc connected to ground with a bias choke and a bias resistor R_{bias}. The LC resonator at the gate consists of an inductor L_r with a quality factor (Q) of 20 at 4 GHz and a varactor diode. The varactor diode was realized using a depletion FET with the drain and source connected together. The oscillation frequency can be controlled by the applied tuning voltage V_{varactor}. Since the gate is connected to ground, the tuning voltage is positive. A varactor with a periphery of 1000 μm was used, which provides a good tradeoff between size, Q, and the tuning range of the resonator. The diode characteristics of this varactor were extracted from S-parameter measurements and are shown in Fig. 4. The output capacitance of the FET, which is determined through the gatewidth (w_a) of the FET, was used as the class-E capacitance C_s. A negative feedback ($\mathrm{Re}\{Z_i\}$) of around $-200\,\Omega$ was chosen to ensure oscillation over the whole band of interest.

The elements of the circuit were optimized with PSPICE using the idealized calculations as starting values, which have been described in Section II. The design goal was to reach maximum efficiency. A gradient optimizer was well suited for this purpose. The FET was modeled with a TOM 2 large-signal model [7]. Previous simulations showed that it was very difficult to optimize the circuit without appropriate starting values because there are many free parameters and local maximums. The validity of the equivalent circuits is limited, especially for the class-E output circuitry. Resistive losses, the finite dc feed, the nonideal switching, the source feedback, and the weak filtering of the harmonics limit the accuracy of the equivalent class-E output circuit. However, the equivalent circuits were well suited to finding starting values

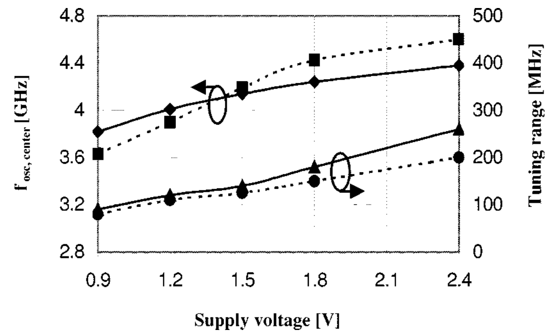

Fig. 7. Oscillation center frequency and tuning range. Solid line: simulated, dotted line: measured.

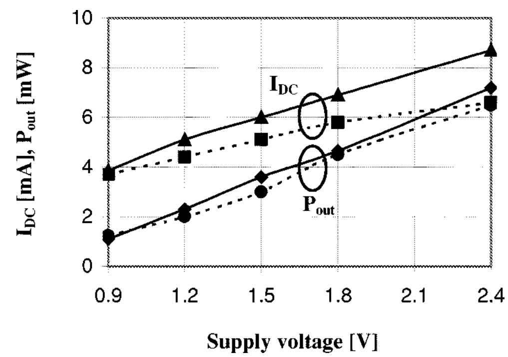

Fig. 8. Output-power and dc current versus supply voltage. Solid line: simulated, dotted line: measured.

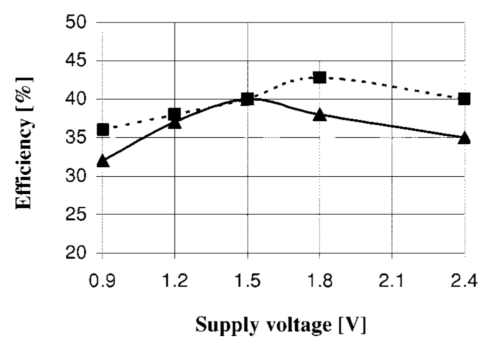

Fig. 9. Efficiency versus supply voltage. Solid line: simulated, dotted line: measured.

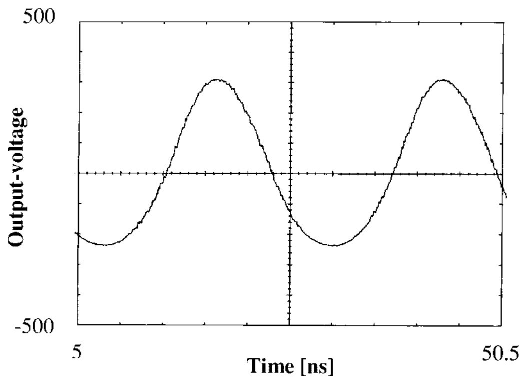

Fig. 10. Measured output-voltage waveform $V_{\rm supply} = 1.2$ V, $V_{\rm varactor} = 0.6$ V.

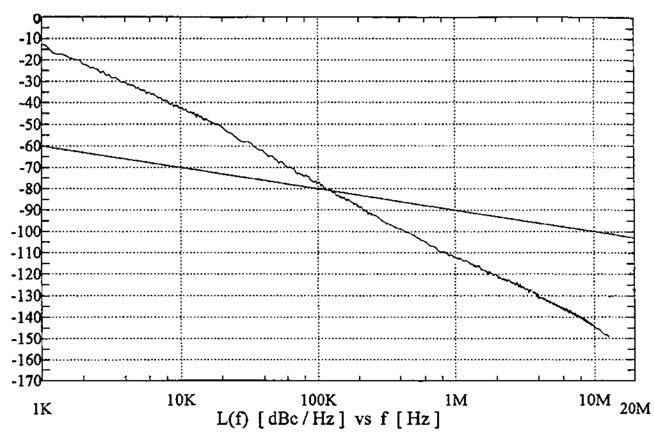

Fig. 11. Phase noise versus frequency offset $V_{\rm supply} = 1.2$ V, $V_{\rm varactor} = 0.6$ V.

for the large-signal optimization. Table II shows a comparison between the calculated and optimized values at 4 GHz with a supply voltage of 1.2 V and a tuning voltage of 0.6 V. The deviations between the calculated and optimized values are relatively small.

Fig. 5 shows the simulated class-E waveforms of the optimized circuit at a supply voltage of 1.2 V and a tunning voltage of 0.6 V. V_S is the effective drain–source voltage of the FET and I_S is the current through the FET. V_S and I_S show typical class-E operation.

IV. Results

Fig. 6 shows a microphotograph of the class-E VCO monolithic microwave integrated circuit (MMIC), bonded into a test package. The chip size is 0.9 mm $\times$ 0.9 mm.

Fig. 7 shows the simulated and measured oscillation frequency and frequency tuning range versus the supply voltage. The tuning voltage was varied from zero to $V_{\rm supply}$. At a supply voltage of 1.8 V, the measured oscillation center frequency is 4.4 GHz. A frequency tuning range of 150 MHz was measured. With an applied supply voltage of 0.9 V, the measured oscillation center frequency was 3.6 GHz with a frequency tuning range of 80 MHz.

Figs. 8 and 9 show the simulated and measured output power, dc current, and efficiency versus the supply voltage. A calibrated HP-436A power meter was used for the power measurements. The losses of the test package were deembedded. An efficiency of up to 43% and an output power of 6.5 dBm were measured with a supply voltage of 1.8 V. At a supply voltage of 0.9 V, an efficiency of up to 36% and an output power of 1.1 dBm were measured. The efficiency of the VCO falls with supply voltages above 1.8 V because the circuit was optimized for 1.2 V, and the class-E mismatch increases with increased supply voltage.

Fig. 10 shows the output-voltage waveform of the VCO at a supply voltage of 1.2 V, measured with a Tektronix 11801 sampling oscilloscope. The harmonics were measured with an HP-8563E spectrum analyzer. The suppression of the first and the second harmonic were -16 and -26 dBc, respectively, at a supply voltage of 1.2 V. This suppression is sufficient for most oscillator applications, thus no additional filtering of harmonics at the output is needed. The phase noise was measured with a HP-E5500 phase noise analyzer using the delay-line method.

Fig. 11 shows the measured phase noise versus the frequency offset for a supply voltage of 1.2 V and a varactor voltage of 0.6 V. The phase noise is -132 dBc/Hz at a frequency offset of 5 MHz, which is sufficient for most applications. We have to note that the oscillator was optimized for low power supply and not for low phase noise.

V. Conclusion

This paper has demonstrated the design and results of a class-E VCO MMIC operating with ultra-low supply voltages. Large-signal optimization was performed using starting values, which were analytically calculated by simple equivalent circuits.

It has been shown that even at supply voltages down to 0.9 V, high-efficiency oscillators can be designed using a standard 0.6-μm GaAs MESFET process. The proposed VCO contributes considerably to the reduction of power consumption and battery size of receivers.

References

[1] J. J. Kucera and U. Lott, "Low DC power cascode LNA's for 1.6 GHz and 5.2 GHz wireless applications," in *European Solid State Circuits Conf.*, Sept. 1998, pp. 336–338.

[2] E. W. Bryerton, W. A. Shiroma, and B. Popovic, "A 5 GHz high efficiency class-E oscillator," *IEEE Microwave Guided Wave Lett.*, vol. 6, pp. 441–443, Dec. 1996.

[3] M. Prigent, M. Camiade, G. Pataut, D. Reffet, J. Nebus, and J. Obregon, "High efficient free running class-F oscillator," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1995, pp. 1317–1320.

[4] A. Warren, J. M. Golio, and W. L. Seely, "Large and small signal oscillator analysis," *Microwave J.*, pp. 229–246, May 1989.

[5] Lee, S. Nam, Y. Kwon, and K. Yeom, "Analytic design of high efficiency harmonic loading oscillator using harmonic two signal method," in *IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 3, 1997, pp. 1495–1498.

[6] F. H. Raab, "Idealized operation of the class-E tuned power amplifier," *IEEE Trans. Circuits Syst.*, vol. CAS-24, pp. 725–735, Dec. 1977.

[7] D. H. Smith, "An improved model for GaAs MESFETs," Triquint, Hillsboro, OR, Tech. Rep., Feb. 1995.