Umut GUVENC et al. / IU-JEEE Vol. 16(1), (2016), 1965-1970

ZERO-IF SECOND HARMONIC SIGE MIXER WITH DC OFFSET CANCELLATION

Umut GUVENC[1], Osman PALAMUTCUOGULLARI[2], Siddik YARMAN[3]

[1]National Research Institute of Electronics and Cryptology, TUBITAK, Turkey
[2]Faculty of Electrical and Electronic Engineering, Istanbul Technical University, Turkey
[3]Department of Electrical and Electronics Engineering, Istanbul University, Turkey
umut.guvenc@tubitak.gov.tr, opal@itu.edu.tr, yarman@istanbul.edu.tr

Abstract:*A new second harmonically pumped SiGe mixer circuit topology is proposed. The proposed circuit is developed to be employed in integrated circuit receivers which use Zero-IF mixers. The proposed circuit topology is analyzed to determine the appropriate operating conditions. It has a unique mechanism to prevent the local oscillator (LO) leakage to radio frequency (RF) signal which results in DC offset voltage at the output. Detailed analysis results are given regarding the biasing dependence of the conversion voltage gain and the linearity of the produced pumping current. The results are also verified by simulation results. Simulation results show that the DC offset voltage at the output stays more than 80 dB below the intermediate frequency (IF) signal.*

Keywords:*DC offset, direct conversion, SiGe mixer, sub-harmonic mixer, Zero-IF.*

1. Introduction

In the current literature, the use of Zero-IF mixer circuits in radio frequency integrated circuit (RFIC) receivers is generally preferred to their superheterodyn counterparts [1], [2]. Various researchers are still striving hard to realize the simple and practically realizable Zero-IF mixer structures [1], [2]. It is known that this type of mixing has serious inherent problems [1], [2]. The most challenging problem is called DC Offset voltage. It is the result of local oscillator (LO) leakage to the input and then mixing with the LO signal itself inside the mixer.

In a second harmonically pumped Zero-IF mixer, the LO frequency is the half of the radio frequency (RF) signal carrier frequency. As it is explained in Section III, the fundamental term of the mixing product is suppressed and only the second harmonic term of this product is taken as the baseband signal [3]. Since the fundamental LO frequency is half of the carrier frequency, the DC offset voltage problem due to the fundamental LO signal leakage is prevented naturally.

To clarify the technical problem, the literature is reviewed for the SiGe sub-harmonic mixers and the solutions proposed to the DC offset problem. Recent works [4]–[6] utilizing the advantages of SiGe BiCMOS technology show the state of the art. Various approaches are proposed for the cancellation of the DC offset voltage by even suppressing the LO leakage or using digital signal processing techniques in [7]–[9].

In the proposed mixer topology, by using second harmonic mixing technique, the reason of the DC offset voltage is reduced to the leakage of the second harmonic of the LO signal. Especially in integrated receiver solutions the leakage of the second harmonic of the LO signal through the substrate and the supply lines may still cause considerable DC offset voltage. The proposed topology contains a unique mechanism to cancel the DC offset voltage by suppressing the leakage of the LO signal.

Upon a unique biasing circuitry, a constant total DC current flows through the overall mixer circuit. First the constant DC current is split into two and the pumping current containing the proper frequency components is obtained.

After performing the mixing, the split pumping current and the residue current is combined again to form the constant DC current. Thus, the total current drawn from the supply voltage is kept constant. Thanks to this attribute of the proposed mixer topology, the noisy large signal mixing operation take place locally in the mixer circuit and it is isolated from the supply lines. The locally produced pumping current and the residue current contain the same frequency components with opposite signs. The leakage from the pumping current to the RF signal path is suppressed by the contrary leakage from the residue current through the symmetric structure.

2. The Proposed Mixer Topology

The proposed second harmonically pumped mixer topology is depicted in Fig. 1. The balanced input RF signal $v_{RF}(t) = V_{RF}\cos\omega_{RF}t$ is applied to the input ports of the both primary and secondary differential pairs and the balanced LO signal $v_{LO}(t)$ is applied to the inputs of the tail transistors Q5 and Q6 which act as the current sources of the primary differential pair. The differential LO signal $v_{LO}(t)$ is defined as

Received on: 04.12.2015
Accepted on: 30.12.2015

$$v_{LO}(t) = v_{LO}^{+}(t) - v_{LO}^{-}(t)$$
$$v_{LO}(t) = (V_C + v_0(t)) - (V_C - v_0(t)) \qquad (1)$$

where $v_0(t) = V_0 \cos\omega_0 t$ is the sinusoidal part and V_C is the DC voltage level of the LO. Transistor Q7 which act as the current source of the secondary differential pair is biased with constant V_B voltage.

Together with the Q5 and Q6 transistors, transistor Q7 compose a differential pair like structure with three transistors in the LO stage. This three transistor differential pair is driven by a constant current I_T. The biasing current of the second differential pair, i_R, is formed as the residue current due to the three transistor differential pair structure. It is the remaining part of the subtraction of i_C from the total biasing current I_T of the LO stage. The expressions of the the biasing currents i_C and i_R are given in Section 3. By arranging the DC biasing levels the LO stage properly, the circuit may be adjusted for the second harmonically pumped mixing. The proper biasing conditions and the effects of the biasing on the conversion voltage gain are also examined in Section 3.

In the proposed topology, the primary differential pair at the RF stage performs the actual second harmonic mixing upon proper biasing. The downconverted baseband signal$v_{IF}(t) = V_{IF}\cos\omega_{IF}(t)$, which is conventionally called the intermediate frequency (IF) signal, occurs on the load impedance Z_L.

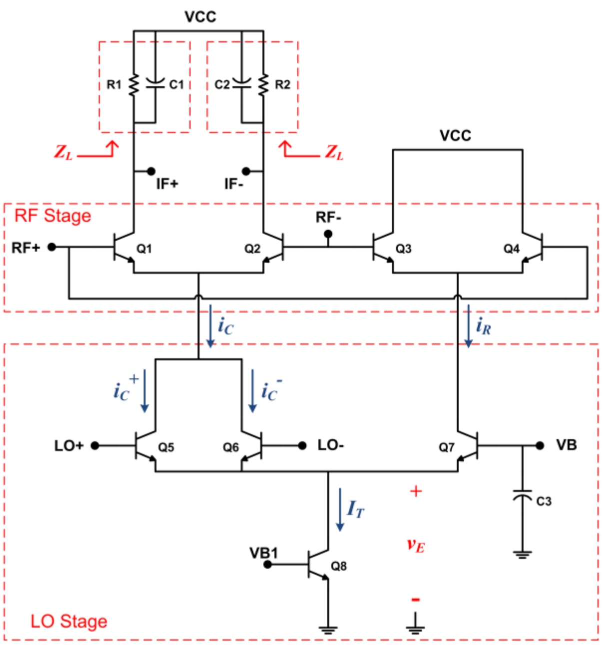

Figure 1. Schematic of the proposed mixer

The secondary differential pair is placed as a dummy structure which is an identical copy of the primary differential pair. Considering that the sum of the tail currents $i_C + i_R$ is constant, when expanded to Fourier series it can be shown that the residue current i_R includes exactly the same frequency components with opposite signs with i_C. For all the frequency components of the pumping current i_C leaking to RF input, there exist an opposite signed leakage due to the residue current i_R and the dummy secondary differential pair. Therefore, the leakage of all the frequency components including the dominant second harmonic frequency of the LO signal, which may result in DC Offset voltage at the output is suppressed. This is the main attribute of the proposed topology.

3. Circuit Analysis of the Proposed Topology

The expressions of i_C and i_R currents can be obtained from the large signal analysis of the three transistor differential pair at the LO stage. By using the expressions for the input voltages, the total biasing current I_T can be written in terms of transistor collector currents as

$$I_T = I_S e^{\frac{-v_E}{V_T}}\left(e^{\frac{v_{LO}^{+}(t)}{V_T}} + e^{\frac{v_{LO}^{-}(t)}{V_T}} + e^{\frac{V_B}{V_T}}\right) \qquad (2)$$

By using the expression for $I_S e^{\frac{-v_E}{V_T}}$ derived from Eqn (2) and performing appropriate trigonometric conversion, the expressions for i_C and i_R can be obtained as

$$i_C = I_S e^{\frac{-v_E}{V_T}}\left(e^{\frac{v_{LO}^{+}(t)}{V_T}} + e^{\frac{v_{LO}^{-}(t)}{V_T}}\right)$$
$$i_C = \frac{I_T}{1+\frac{A}{2}\operatorname{sech}\left(\frac{1}{V_T}v_0(t)\right)} \qquad (3)$$

$$i_R = I_S e^{\frac{-v_E}{V_T}} e^{\frac{V_B}{V_T}}$$
$$i_R = \frac{I_T}{1+\frac{2}{A}\cosh\left(\frac{1}{V_T}v_0(t)\right)} \qquad (4)$$

where A is a constant defined by the DC biasing conditions of the LO stage. The definition of A is given as

$$A = e^{\frac{V_B - V_C}{V_T}} \qquad (5)$$

When investigated by expanding to Fourier series, it is approved that both i_C and i_R contain only the even harmonics and i_R contains same but oppositely signed terms with i_C except the DC term.

Let a transfer function $F_C(x)$ is defined for the transformation from the sinusoidal term of the LO signal $v_0(t)$ to pumping current i_C as

$$F_C(x) = \frac{I_T}{1+\frac{A}{2}\operatorname{sech}\left(\frac{1}{V_T}x\right)} \qquad (6)$$

The graph of $F_C(x)$ is given in Fig. 2 along with applied $v_0(t)$ and resulting $i_C(t)$ to illustrate the conversion from $v_0(t)$ to $i_C(t)$. From Fig. 2, the effects of the biasing dependent variables I_T and A on $i_C(t)$ can be observed.

The pumping current i_C varies the transconductance of the primary differential pair accordingly. The expression of the transconductance g_m of the transistors Q1 and Q2 can be written as

$$g_m = \frac{1}{2V_T}\frac{I_T}{\left(1+\frac{A}{2}\operatorname{sech}\left(\frac{1}{V_T}v_0(t)\right)\right)} \tag{7}$$

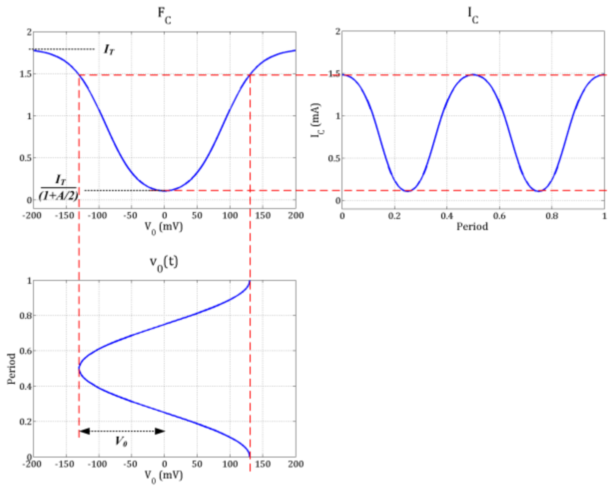

Figure 2. Conversion from $\boldsymbol{v_0(t)}$ to $\mathbf{i_C}(\boldsymbol{t})$ via transfer function $\boldsymbol{F_C}$

Since g_m expression given in Eqn (7) is an even periodic function, the Fourier series expansion given in Eqn (8) of which coefficients are as in Eqn (9) can be used.

$$g_m = g_{m0} + 2\sum_{n=1}^{\infty} g_{mn}\cos n\omega_0 t \tag{8}$$

$$g_{mn} = \frac{1}{2\pi}\int_{-\pi}^{\pi} g_m \cos n\theta \, d\theta \tag{9}$$

When Eqn (7) is used in Eqn (9), the expression for the coefficients of the transconductance g_m is obtained as

$$g_{mn} = \frac{I_T}{4\pi V_T}\int_{-\pi}^{\pi}\frac{\cos n\theta}{1+\frac{A}{2}\operatorname{sech}\left(\frac{V_0}{V_T}\cos\theta\right)}d\theta \tag{10}$$

where θ is used for $\omega_0 t$. Evaluating the integrals for the coefficients, it is inspected that the transconductance is mainly composed of the even harmonics of the LO. The elimination of the odd harmonics from the total transconductance variation of the mixer prevents fundamental mixing products at the output.

The output IF current $i_{IF}(t)$ is produced upon the multiplication of the input RF signal $v_{RF}(t) = V_{RF}\cos\omega_{RF} t$and the transconductance g_m. The output IF voltage $v_{IF}\ (t)$ occurs on the load impedance Z_L. Because of the low pass characteristic of the load impedance Z_L, all the higher frequency terms of $v_{IF}\ (t)$is filtered out and only the IF term at the baseband exists. At the baseband frequency, the load impedance Z_L can be simplified into only the real resistive part of it R_L. Since the IF term at the output is related to the second harmonic of the LO frequency, the expression of $v_{IF}\ (t)$ can be simplified into Eqn (12).

$$i_{IF}(t) = v_{RF}(t)\left(g_{m0} + 2\sum_{n=1}^{\infty} g_{mn}\cos n\omega_0 t\right) \tag{11}$$

$$v_{IF}(t) = g_{m2}\, V_{RF} R_L \cos\omega_{IF} t \tag{12}$$

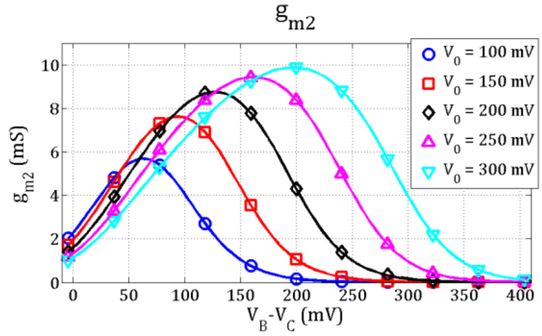

Figure 3. The value of $\boldsymbol{g_{m2}}$ according to $(\boldsymbol{V_B - V_C})$ for different values of $\boldsymbol{V_0}$

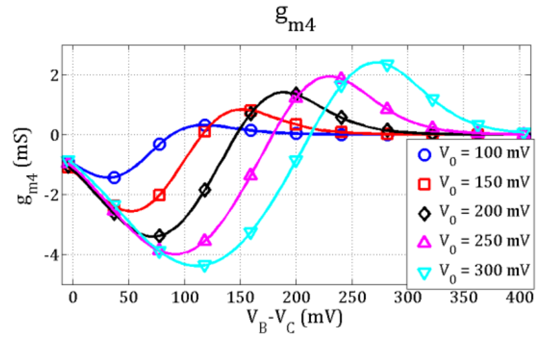

Figure 4. The value of $\boldsymbol{g_{m4}}$ according to $(\boldsymbol{V_B - V_C})$ for different values of $\boldsymbol{V_0}$

The conversion voltage gain A_{VC} can be expressed as

$$A_{VC} = \frac{v_{IF}}{v_{RF}} = g_{m2} R_L \tag{13}$$

According to the expression in Eqn (13), the conversion voltage gain of the circuit is directly determined by the second harmonic coefficient g_{m2} of the Fourier expansion of g_m. The equation for g_{m2}, due to the definition of A given in Eqn (5), depends on the voltage difference $(V_B - V_C)$ and the magnitude of LO signal V_0. The parametric graph of the g_{m2} variation according to $(V_B - V_C)$is shown in Fig. 3. According to the graph, the value of g_{m2} can be maximized by increasing the V_0 and finding the appropriate voltage difference $(V_B - V_C)$.

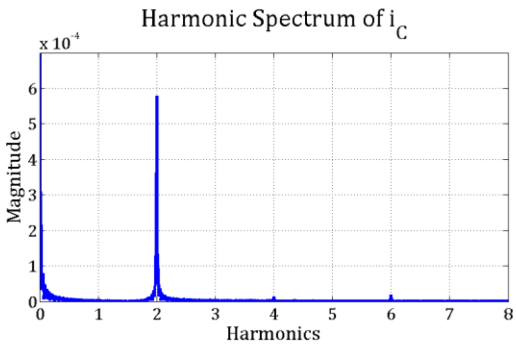

Figure 5. The harmonic spectrum of $\boldsymbol{i_C}$ when the value of $(\boldsymbol{V_B - V_C})$ is optimized for $\boldsymbol{V_0 = 110mV}$

Besides the dependence of the second harmonic coefficient g_{m2} on $(V_B - V_C)$ and V_0, the linearity of the resultant pumping current i_C is also strictly determined by the values of voltage difference $(V_B - V_C)$ and V_0. Fig. 4 shows the forth harmonic coefficient g_{m4} which is the largest coefficient after g_{m2}. It is observed that at a specific biasing point for both $(V_B - V_C)$ and V_0, g_{m4} passes from zero, where at the same biasing point g_{m2} is almost at its maximum.

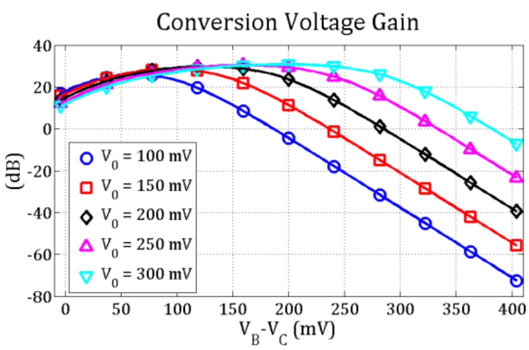

Figure 6. The conversion voltage gain $\boldsymbol{A_{VC}}$ according to $(\boldsymbol{V_B - V_C})$ for different values of $\boldsymbol{V_0}$

Fig. 5 shows the harmonic spectrum obtained from the Fourier transformation of the pumping current i_C when $(V_B - V_C)$is set to optimum value for $V_0 = 110mV$. The harmonic spectrum verifies that under the specified conditions the $i_C$is linear and mainly consists of the second harmonic of the LO.

By using the obtained g_{m2} values in the conversion voltage gain expression given in Eqn (13), the graph in Fig. 6 can be plotted for the same sweep range of voltage difference $(V_B - V_C)$ and the same values of parameter V_0.

4. Simulation Results

The proposed second harmonic mixer circuit was created and appropriate simulations were performed on the circuit schematic in AWR Analog Office design environment. A 0.18μm SiGe BiCMOS processing technology was used for the simulations of the circuit.

Harmonic balance simulations were performed on the circuit schematic for exactly the same biasing conditions as it is in the circuit analysis section for the comparison of the simulation and analysis results. For the harmonic balance simulations, the biasing voltage difference $(V_B - V_C)$ was swept from 0 V to 400 mV. The LO signal amplitude V_0 was changed from 100 mV to 300 mV with 50 mV steps as a parameter. The RF signal frequency was selected to be 2 GHz and the LO frequency was set as 0.995 GHz to obtain an IF signal frequency of 10 MHz at the output. The resultant voltage conversion gain A_{VC} graph is given in Fig. 7. and the harmonic spectrum of the pumping current i_C is given in Fig. 8. The simulation results show that the proposed mixer topology can supply more than 20 dB conversion voltage gain. Upon proper biasing for a given LO amplitude, the conversion voltage gain can be maximized.

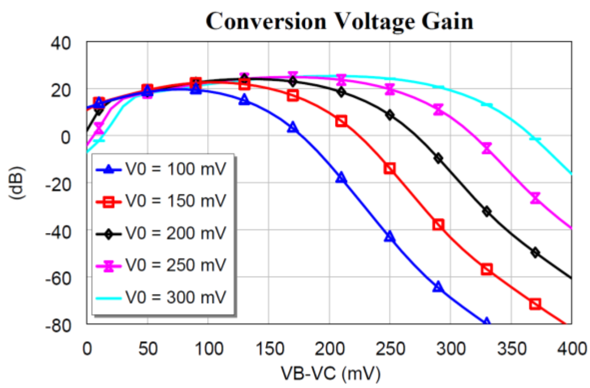

Figure 7. The conversion voltage gain A_{VC} obtained from the simulation results

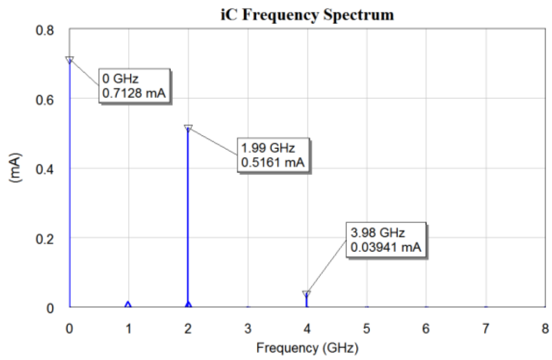

Figure 8. The harmonic spectrum of i_C obtained from harmonic balance simulation for the optimum value of $(V_B - V_C)$ for $V_0 = 110mV$

For further investigation of the circuit performance, harmonic balance simulation was performed with RF frequency swept from 2 GHz to 10 GHz. In this simulation, LO frequency was changed accordingly to keep the output IF signal frequency constant at 10 MHz. The resultant conversion voltage gain graphic given in Fig. 9 proves the applicability of the topology for wide band applications.

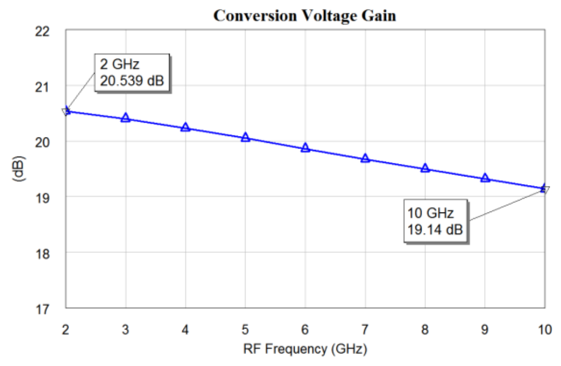

Figure 9. The change of conversion voltage gain upon RF frequency sweep from 2 GHz to 10 GHz

From the harmonic balance simulations performed on the circuit schematic it is also observed that the DC offset voltage at the mixer output stays more than 80 dB below the IF signal component. This amount of high DC offset voltage suppression indicates the effectiveness of the proposed topology as the Zero-IF mixer.

5. Conclusions

A second harmonically pumped Zero-IF mixer topology which is realized with SiGe technology is proposed. The proposed mixer topology is appropriate for being used within a wide frequency range. The circuit analysis of the proposed mixer topology is performed and the operating conditions are examined. It is exhibited that the mixer topology performs second harmonic mixing while the fundamental mixing products are suppressed precisely at the output. It is also shown that by optimizing the biasing conditions, the conversion voltage gain can be maximized. Thanks to the unique mechanism employed in the topology, the leakage from LO to RF is suppressed. Thus, the DC offset voltage problem at the IF output is prevented. It is shown that the dependency of the conversion voltage gain to the biasing conditions in simulations is in agreement with the analysis results. Conversion voltage gain more than 20 dB and a DC offset voltage suppression more than 80 dB is obtained.

6. References

[1] H. M. Hsu and T. H. Lee, “A Zero-IF Sub-Harmonic Mixer with High LO-RF Isolation Using 0.18 µm CMOS Technology,” in Proceedings of the 1st European Microwave Integrated Circuits Conference, Manchester, UK, September 2006, pp. 336–339.

[2] S. A. S. Mohamed and Y. Manoli, “Design of Low-Power Direct Conversion RF Front-End with a Double Balanced Current-Driven Subharmonic Mixer in 0.13 µm CMOS,” IEEE Transactions on Circuits and Systems, vol. 60, no. 5, pp. 1322–1330, May 2013.

[3] E. Degenford, M. Cohn, and A. Newman, “Harmonic Mixing with an Anti-Parallel Diode Pair,” IEEE Transactions Microwave Theory and Techniques, vol. 23, no. 8, pp. 667–673, August 1975.

[4] L. Sheng, J. Jensen, and L. Larson, “A Wide-Bandwidth Si/SiGe HBT Direct Conversion Sub-Harmonic Mixer/Downconverter,” IEEE Journal of Solid-State Circuits, vol. 35, no. 9, pp. 1329–1337, September 2000.

[5] X. Yang, Z. Sun, T. Shibata, and T. Yoshimasu, “W-Band Low Power Sub-Harmonic Mixer IC in 130-nm SiGe BiCMOS,” in IEEE International Wireless Symposium, Shenzhen, March 2015, pp. 1–4.

[6] J.-S. Syu and C. Meng, “15 GHz High-Isolation Sub-Harmonic Mixer with Delay Compensation,” IEEE Microwave and Wireless Components Letters, vol. 19, no. 12, pp. 810–812, December 2009.

[7] B. Matinpour, S. Chakraborty, and J. Laskar, “Novel DC-Offset Cancellation Techniques for Even-Harmonic Direct Conversion Receivers,” IEEE Transactions on Microwave Theory and Techniques, vol. 48, no. 12, pp. 2554–2559, December 2000.

[8] C. de la Morena-Alvarez-Palencia, K. Mabrouk, B. Huyart, and et al., “Direct Baseband I-Q Regeneration Method for Five-Port Receivers Improving DC-Offset and Second-Order Intermodulation Distortion Rejection,” IEEE Transactions on Microwave Theory and Techniques, vol. 60, no. 8, pp. 2634–2643, June 2012.

[9] S. Yamada, O. Boric-Lubecke, and V. M. Lubecke, “Cancellation Techniques for LO Leakage and DC Offset in Direct Conversion Systems,” in IEEE MTT-S International Microwave Symposium Digest, Atlanta, GA, USA, June 2008, pp. 1191–1194.

Note:

Umut Guvenc was born in Turkey in 1980. He received his B.Sc. and M.Sc. degrees in electronic engineering from Istanbul Technical University, Turkey, in 2003 and 2006. He attended Ph.D. program at Istanbul Technical University in 2006 in the area of zero-IF receivers and second harmonic mixers under the supervision of Prof. Osman Palamutcuogullari. In 2005, he joined Semiconductor Research Laboratory (YITAL) at Scientific and Technological Research Council of Turkey (TUBITAK) as a researcher responsible for application specific integrated circuit (ASIC) design.

Osman Palamutcuogullari received his B.Sc. and M.Sc. degrees from the Electrical Engineering Department of Istanbul Technical University, in 1968 and 1971, respectively. He received Ph.D. degree from the Electronics and Communication Department of Bradford University, England, in 1974. He became associate Professor on Electronics Circuit Elements and Devices in1982 and became Professor on Microwave and RF Engineering in 1988. Osman Palamutcuogullari has presented more than 80 published scientific papers in different conferences and congress at different times. He has so far lectured on

Electronics Circuits and Systems, Communication Electronics Circuits, Microwave and RF Engineering, Communication Systems.

Siddik Yarman (M'82--SM'88--F'04) received the B.Sc. degree in electrical engineering from the Technical University of Istanbul, Istanbul, Turkey, in 1974, the M.Sc. degree from the Stevens Institute of Technology, Hoboken, NJ, in 1978, and the Ph.D. degree from Cornell University, Ithaca, NY, in 1982. He was a Member of Technical Staff with the Microwave Technology Center, David Sarnoff Research Center, Princeton, NJ. He was a Professor with Anatolia University, Eskisehir, Turkey, Middle East Technical University, Ankara, Turkey, the Technical University of Istanbul, Istanbul, Turkey, and Istanbul University, Istanbul, Turkey. He was one of the founders of I-ERDEC Maryland USA (1983), STFA SAVRONIK (a defense electronics company in Turkey) (1986), and ARES Security Systems Inc. (1990). He was the Chief Technical Adviser to the Turkish Prime Ministry Office and the Director of Electronic and Technical Security of Turkey (1993--1999). He was the founding President of Isik University, Istanbul, Turkey (1996--2004). He was a Visiting Professor with Ruhr University, Bochum, Germany (1987--1994) and the Tokyo Institute of Technology, Tokyo, Japan (2006--2008). He has authored or coauthored over 200 scientific and technical papers in the field of electrical/electronic engineering, microwave engineering, computer engineering, mathematics, and management. He holds four U.S. patents assigned to the U.S. Air Force. Dr. Yarman has served on various technical and scientific committees since 1980 in the U.S. and Turkey. He is an Alexander Von Humboldt Research Fellow (1987), Bonn, Germany, and a member of the New York Academy of Science (1994). He was the recipient of the Young Turkish Scientist Award (1986), the Technology Award of the National Research and Technology Council of Turkey (1987), and Man of the Year in Science and Technology, Cambridge Biography Center, Cambridge, U.K. (1988).