# A Low Noise Vector Modulator with integrated Basebandfilter in 120 nm CMOS Technology

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***Abstract*** **— A low noise vector modulator for cellular systems with integrated baseband filter has been developed in 120 nm CMOS technology. The baseband filter is an anti-aliasing filter for the digital to analogue converter of the baseband that has been build around a differential amplifier with 3rd order butterworth low pass characteristic. High linearity and low output noise can be achieved for the modulator due to the current interface between the baseband filter and the mixer cells. The GSM specification for emitted noise into the receive band can be fulfilled without using a duplex filter behind the power amplifier output.**

## I. Introduction

Single chip transceiver for cellular systems have been developed in standard CMOS technology with a performance that is comparable to bipolar circuits regarding power consumption and chip size [1]. Manufacturing costs of RF CMOS technology are low and it is possible to integrate the analogue RF circuits into the digital baseband for single chip solutions. But the costs of the whole radio system remain high if external components like preamplifier and discrete filters have to be added at the terminals of the transceiver. It is therefore necessary to achieve low noise and high power at the output of the modulator to avoid SAW filters and preamplifier for the power amplifier in the transmit chain. Vector modulation is the preferred transmitter architecture for future GSM EDGE and UMTS networks where linear modulation schemes with amplitude modulation like 8-PSK with 3/8π phase rotation and QPSK are employed for high data rates. IF-VCOs and IF-filters can be avoided when a direct conversion modulator is used for the transmitter.

## II. Transmitter Architecture

Fig. 1 shows the block diagram of a GSM transmitter with vectormodulation. The vector modulator chip consists of differential baseband filters with current outputs, frequency divider and two mixers cells for each frequency band with differential inputs. The quadrature LO signals with frequency $\omega_C$ are generated by the divider that is build up with standard Master Slave Flip Flops. The impact of VCO load pulling is low, because the VCO frequency is four times the transmit frequency. The mixer cells of the modulator are connected via a current interface to the baseband filter. The output signal of the modulator RF(t) is the sum of the two mixer output signals that are modulated with the modulation frequency $\omega_m$. It is given by:

$$RF(t) = G \cdot \begin{bmatrix} \cos(\omega_m t)\cos(\omega_c t + \Phi) + D \cdot \cos(\omega_c t) \\ -\sin(\omega_c t) \cdot \cos(\omega_m t + 90°) \end{bmatrix} \quad (1)$$

Where G is the gain of the modulator. The unwanted lower sideband is created by the phase error Φ of the quadrature LO. The amplitude of the carrier feedthrough is given by D. A SAW filter between the modulator output and the power amplifier filters noise and spurious emissions.

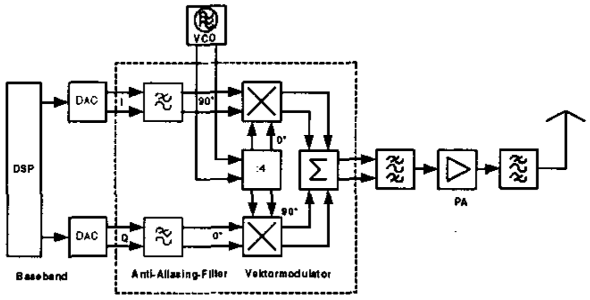


Fig. 1. Blockdiagram of vector modulator

### *A. Baseband filter*

The quadrature baseband currents for the mixer inputs are generated by a voltage controlled current source that is built up around a differential operational amplifier with resistive and capacitive feedback and two PMOS transistors as shown in Fig. 2. The filter transfer function is a third order butterworth filter and is given by:

$$\Delta I_D = \frac{\Delta U_{in}}{2R_D} \cdot \frac{A}{\left(1 + k_1 \cdot \frac{j\omega_m}{\omega_{3dB}} + k_2 \cdot \left(\frac{j\omega_m}{\omega_{3dB}}\right)^2 + k_3 \cdot \left(\frac{j\omega_m}{\omega_{3dB}}\right)^3\right)} \tag{2}$$

The DC gain A is determined by the resistor feedback of the operational amplifier. The coefficients k1, k2, k3 are the butterworth filter parameter determined by the feedback of the operational amplifier. The filter serves as an anti aliasing filter for a baseband DAC with 13 MHz sampling frequency and has a filter cut-off frequency of 450 kHz, that is sufficient wide for the GMSK modulation. The drain current through the PMOS transistors is controlled by the output voltage of the operational amplifier. The bias current of the PMOS transistors can be adjusted with the common mode control voltage of the operational amplifier.

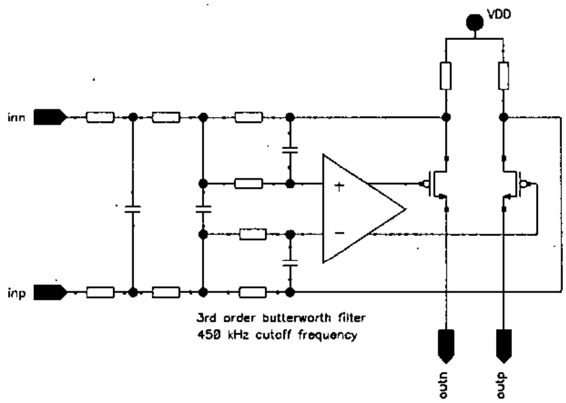


Fig. 2. Basebandfilter with current output

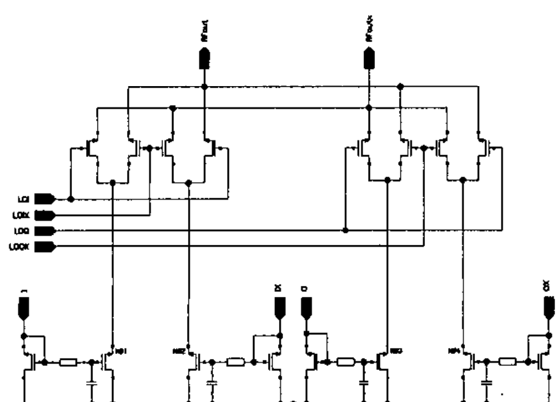

Fig. 3. Schematic of vector modulator

Using baseband filters with current output that directly fed the mixer avoids the need of a differential amplifier with resistive feedback for voltage to current conversion in the mixer cell. A differential amplifier would reduce the linearity and increase the noise. Therefore a high linearity modulator can be realized with low bias current with the proposed configuration.

*B. Mixer*

A current folded mixer structure as shown in Fig. 3 has been chosen for low supply voltage [2]. The supply voltage of the RF circuits is only 1.5 V to protect the short channel transistors with drain source breakdown voltage of 1.5 V. The quadrature baseband currents are fed into the switching core transistors via current mirrors and amplified with a gain of 14 dB before being mixed up to transmit frequency. The mixer cells are directly supplied with baluns at the drain terminals of the switching RF transistors. The differential balun load of 200 Ω is converted into a single ended 50 Ω impedance. No more amplifier stages are needed behind the mixer that could reduce linearity and create further noise.

## III. Output Noise

A problem of designing radio circuits in CMOS is the high flicker noise in comparison to bipolar technology that is given by :

$$S_f = \frac{k}{W \cdot L} \cdot \frac{1}{f} \tag{3}$$

where k is a process dependant constant, f the frequency and W·L are the transistor dimensions. In upconversion mixer mainly the transistors of the small signal RF input stage contribute flicker noise to the output signal but not the switching transistors in the mixer core with large signal input [3]. Noise nearby the signal carrier increases the phase error of the digitally modulated transmitter signal. The flicker noise can be reduced by increasing the size of the transistors. Increasing transistor dimensions results in higher parasitic capacitances and reduces the transit frequency $f_T$ of the transistors. This problem is essential in receiver design, where the received RF signal has to be amplified with low distortion. But in transmitter design the linear path of the mixer is a low frequency baseband signal where large scale transistors can be used. At higher frequency offsets from the carrier the flicker noise diminishes and the transmitter noise mainly results from phase noise of the VCO and the channel noise of the bias transistors in the mixer cells. The main challenge in GSM transmitter design is to achieve a high signal to noise ratio at the output of the modulator in the transmit band (880-915 MHz) to fulfil the GSM specification for the transmitter noise emitted into the receive-band (925-960 MHz). The noise in the receive-band has to be below -79 dBm at a frequency offset of 20 MHz to the transmit band measured with a spectrum analyser bandwidth of 100 kHz [4]. The highest GSM output power level is Pout = 33 dBm. Hence the output noise at the transmitter output must be at least:

$$S_{20MHz} \le -P_{out} + P_{noiserx} - 10\log(100kHz) \le -162dBc \quad (4)$$

In GSM systems power efficient class C power amplifier are used for the amplification of the phase modulated GMSK signal to achieve long talk time for the mobile station. Because of the PA switching characteristic noise in the transmit band at the output of the modulator is folded into the receive band behind the PA with a typical noise conversion gain of 5 dB. Therefore the signal to noise ratio at the modulator output must be at least -157 dBc to avoid an expensive, bulky duplex filter with high insertion loss behind the output of the PA.

RC filter with a cut-off frequency of 800 kHz have been introduced into the current mirrors of the mixer to reduce the noise generated by the baseband filter. The transistor gate width to length ratio has to be minimized for low transconductance $g_m$ and low channel noise of the bias transistors that is given by:

$$\overline{i_d^2} = 4 \cdot kT\left(\frac{2}{3} g_m\right) \cdot \Delta f \quad (5)$$

A transistor gate length of 600 nm has been chosen for the bias transistors for low flicker noise.

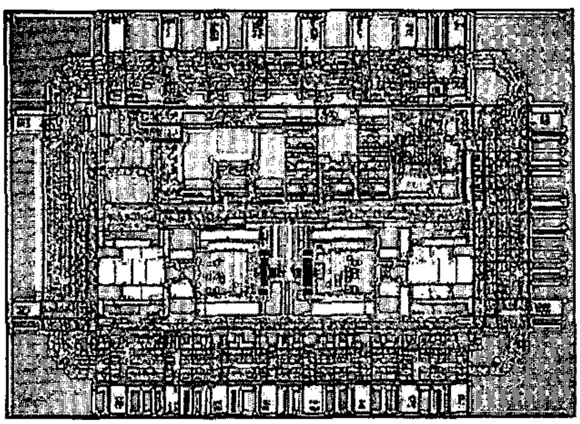

Fig. 4. Chip photograph

## IV. REALIZATION

The vector modulator has been designed in standard 120 nm CMOS technology with six copper metalization layers. The gate oxide thickness is 2.8 nm. Metal-Insulator-Metal capacitances (MIM) of 1 fF/µm² density with low coupling into the substrate are used for the RF-circuits. The drain source voltage of the MOS transistors with 120 nm gate length can be stressed up to 1.5 V. Analogue I/O transistors of 400 nm gate length with 2.5 V breakdown voltage are available that can be used for the design of bandgaps and operational amplifier. The chip size is only 1 mm² without bondpads. The package of the chip is a TSSOP24 housing. The chip photograph is shown in Fig. 4.

## V. MEASUREMENT RESULTS

The vector modulator has been measured with sinusoidal baseband signals. The modulator output power level is 1 dBm for a differential 200 Ω drain load of the balun (see Fig. 5). The output power is sufficient to drive a standard GSM power amplifier with 0 dBm input power. The carrier suppression at the vector modulator output is -29 dBc and mainly results from offset voltages in the baseband filters and the current sources. The image sideband is suppressed by -37 dBc and can be reduced to 40 dBc by adjusting the amplitude mismatch of the baseband signals. This corresponds to a phase error of 1.6°in the quadrature LO signal generation. A third order intermodulation product (IM3) at a frequency offset of four times the modulation frequency $\omega_m$ that can not be suppressed by a filter is -51 dBc below the wanted signal.

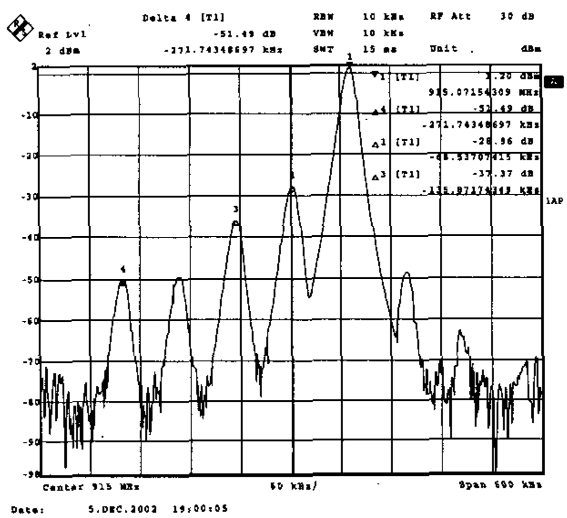


Fig. 5. Modulation spectrum with sinusoidal baseband signals

In Fig. 6 system measurement results with a digital modulated GMSK signal are shown. The margin of the modulated signal in the modulation mask is 10 dB due to the modulators high linearity. The phase error of the GMSK modulated signal is 1.2 ° rms and far below the GSM specification of 5° rms [4]. The dc power consumption of the whole modulator with divider by 4 is only 170 mW. Fig. 7 shows the phase noise measurement of the modulator with the GSM specification for the modulation mask. A signal to noise ratio of 162 dBc/Hz has been achieved at a frequency offset of 20 MHz to the carrier. A duplex filter at the power amplifier output is not needed to fulfil the GSM specification for noise emission into the receive band.

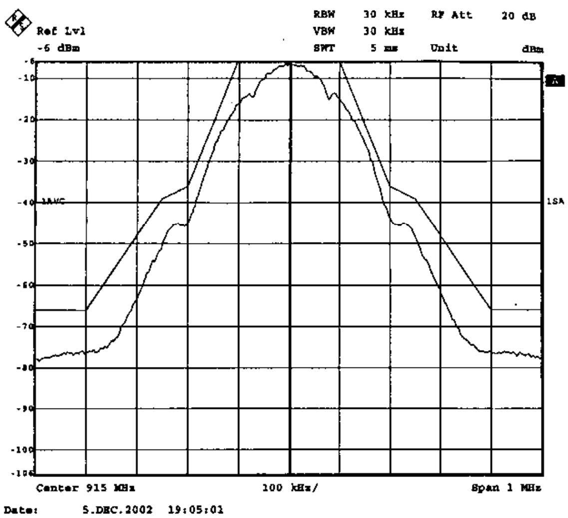


Fig. 6. Modulation spectrum with GMSK modulation

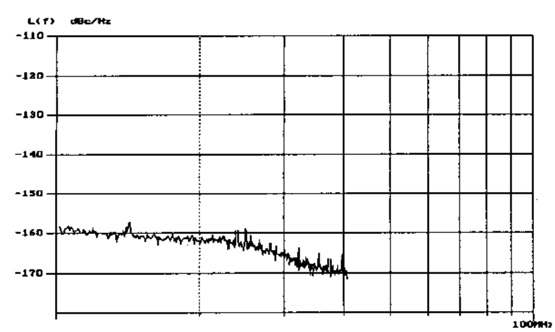


Fig. 7. Transmitter noise emission into the receive band

## VI. Conclusion

A vector modulator designed in 120 nm technology has been presented. The modulator output power is sufficient high to drive a GSM power amplifier. High linearity and low power consumption is achieved by a voltage to current converter for the baseband signals that has been build up around an operational amplifier. The feedback of the converter has 3rd order butterworth lowpass characteristics. The output current is connected to the mixer core via a current mirror. A passive low pass filter in the current mirror suppresses the noise of the baseband circuits. The strict GSM specification for noise emission of the transmitter into the receive band can be fulfilled without needing a duplex filter behind the power amplifier.

## References

[1] M. Borremans, M. Steyaert, "A CMOS 2 V Quadrature Direct Up-Converter Chip for DCS-1800 Integration", *2001 ESSCIRC Proc.*, September 2001.

[2] Tsukahara et al, "A 2 V 2-GHz Si-Bipolar Direct Conversion Quadrature Modulator", *1996 IEEE Journal of Solid State Circuits*, vol. 31, No 2, February 1996.

[3] H. Darabi and A. Abidi, "Noise in RF-CMOS Mixers: A Simple Physical Model," *2000 IEEE Transactions on Solid State Circuits*, vol. 35, pp. 15-25, January 2000.

[4] ETSI 300910 (GSM 05.05 version 5.4.1): Digital cellular telecommunication system (Phase 2+), Radio transmission and reception.