Low cost low power 24 GHz FMCW radar transceiver for indoor presence detection

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***Abstract*—In this paper a first time right 24 GHz FMCW radar transceiver is presented. The MMIC has a low power consumption of 86 mW and an output power of -10 dBm. Due to the integrated IF amplifier, the conversion gain of the receiver is 51 dB and the base band signals are directly processed with an ADC. The developed antenna is a PCB dipole array with 10 dBi gain. Together with a commercial microcontroller this results in a small presence detection sensor required for future intelligent LED lighting systems.**

Keywords—FMCW; radar; SiGe; MMIC; dipole; antenna

I. Introduction

The lighting industry is currently going through a radical transformation, driven by both the rapid progress of Solid State Lighting (SSL) and semiconductor technologies, and the changing societal needs like sustainability and improved energy efficiency [1]. To pursue these immense challenges, research focuses on integration that will enable intelligent LED lighting control systems by using energy-efficient dimming and fast switching capabilities. At system level the integration comprises reliable task and activity sensors, sensing algorithms, a robust architecture and interfaces, essential for smart and intelligent lighting systems. At component level the research focuses on the integration of electronics, controls, LEDs and sensors into a single luminaire. In todays lighting systems a passive infrared (PIR) sensor is commonly used. To detect people at work in an office environment requires a sensor which is sensitive to very small movements. For this kind of detection IR technology can not meet the specifications. The proposed sensor is a FMCW radar, which is able to detect small movements such as a typing motion. In order to embed a sensor into an intelligent luminaire it has to fulfill several requirements. Firstly it has to be low cost, so it will hardly rise the price of a luminaire. Secondly, from the international standards on switching of fixed electronic systems, it is required that the electronics power consumption <500 mW during system standby. Thirdly, in order to achieve an attractive form factor the sensor has to be "paper-thin" and small in size. In this paper the design and characterization of a high frequency, low cost, low power FMCW radar front-end and required antennas are presented. Section II describes the system architecture, which comprises of a microcontroller, an MMIC and two antennas. The MMIC design is discussed in section III. The antenna design and characterization are discussed in section IV. In section V the measurement results of the packaged MMIC are compared with the simulation results, followed by the conclusions.

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NXP semiconductors has supplied the wafer run and packaging for the MMIC.

II. System design

The proposed FMCW radar consist of three parts; a commercial ARM microcontroller, a MMIC transceiver and a dipole antenna. Fig 1 depicts a block diagram of the FMCW radar system. In order to minimize the power consumption the frequency sweep is generated with the DAC, avoiding a dedicated PLL with the associated dissipation. The received signal is directly converted to IF with a double sideband mixer without a low noise pre-amplifier. The maximum detectable range is a function of output power, noise figure, phase noise, detection threshold and target size. The maximum detectable speed is inversely proportional to the center frequency and the sweep repetition period. The center frequency is chosen to be 24 GHz as a trade-off between small antenna size and power consumption of the electronics. The sweep repetition period is then a trade-off between the maximum speed and the maximum ADC clock frequency.

III. Chip design

For the on-chip RF blocks as well as the antenna structure a differential topology is chosen, so lossy balun circuitry is dispensable. This reduction in loss results in a more efficient MMIC. The VCO is based on a Colpitts topology, see Fig 2. The transistors Q1 and Q2 together with the capacitors C1, C2, C3 and C4 form the core of the Colpitts oscillator. Because each branch in the oscillator only generates an AC current for half of the time, the rest of the time this branch does not need a bias current[2,3].

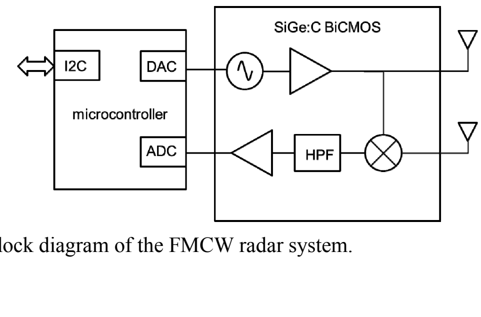

Fig. 1. Block diagram of the FMCW radar system.

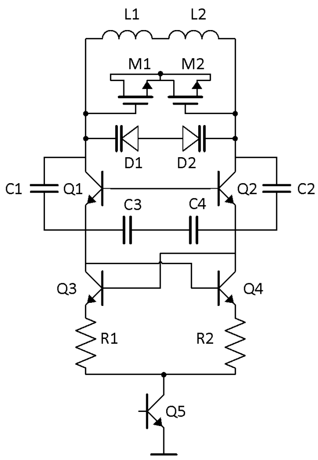

Fig. 2. Schematic of VCO

Therefore the oscillator is biased with a cross coupled negative gm structure (Q3 and Q4) with a single tail current source Q5. The total bias current is reduced by a factor of two, while maintaining the same phase noise performance. The cross coupled structure contains emitter resistances to compensate for asymmetry from non-ideal layout effects, which would otherwise result in the oscillator to fade out. The frequency of oscillation is swept by changing the tune voltage applied from the external DAC circuit to the anode of the two varactors (D1 and D2). To partly compensate a frequency offset due to process variation the gate source capacitance of two NMOS transistors (M1 and M2) is set by connecting the source with an external pull-up or pull-down resistor.

The biasing of the VCO consists of three parts. The main supply is set at 2.7 V by a voltage regulator circuit and is applied through the tank inductor. From this supply a voltage divider is used to set the voltage of the base nodes of the Colpitts transistors (Q1 and Q2). The oscillation frequency dependence on temperature is partly compensated by applying a proportional to absolute temperature (PTAT) reference current, which is mirrored to current source Q5.

A buffer amplifier mitigates the effect of pulling on the frequency of oscillation. This is a two stage amplifier with a differential pair and emitter followers. The emitter followers can delivers an output power of -10 dBm to both the TX output as well as the local oscillator (LO) input of the mixer. The differential pair is required to increase the isolation and prevent pulling due to RF to LO crosstalk of the mixer. The differential amplifier has a parallel resistor and inductor load to optimize the voltage gain. Emitter degeneration has been applied with resistors to improve linearity for the high voltage swings from the VCO. In the emitter follower a series RC network parallel to the input has been used for stabilization. Both amplifier stages are optimized for power consumption, so the collector currents will clip at 0mA when the large voltages from the VCO are applied.

The mixer is designed as a Gilbert-cell mixer topology. The RF input is matched to the differential antenna with a high pass LC network. From a symmetry point of view the network uses two inductors to a virtual common node. Diodes at this common node provide the ESD protection. An RC network at this node suppresses the even mode voltage. At the output of the mixer the IF signal is filtered out by a low pass RC filter. By using a large size for the LO transistors and applying restive feedback in the gm stage a 1-dB compression point of -20 dBm is achieved.

The IF amplifier is a two stage amplifier. The gain of the amplifier is determined with a differential resistive feedback network over the two stages. The feedback network includes a capacitor to ensure stability at high frequencies. The first order HPF is put in front of the IF amplifier and is a simple RC network. The resistance of 200 kΩ is in parallel to the differential input of the amplifier and is used to apply the biasing. To get a corner frequency of 6 kHz, a capacitor of 265 pF is required for each input. Only one of the outputs of the IF amplifier drives the single ended ADC, without a circuit for differential to single ended conversion. An active balun consumes too much power, a passive balun is too large at these low frequencies. A microphotograph of the MMIC is depicted in Fig. 3 and the size of the core without pads is 1.1x0.9 mm^2.

IV. ANTENNA

Development of a small sized radar is here principally facilitated by the choice of frequency. At the same time, other system requirements will affect its size as well. Such, antenna related, requirements are identified as: high efficiency; limited field of view (av. 50°x50° over the frequency band; fixed beam); separate transmit and receive channels; limited cross talk (< -25 dB). As a consequence, the antenna will indeed dominate the mentioned size. A straightforward choice for the antenna type is one that is in line with the differential nature of the amplifier. Application of baluns, requiring considerable space to make them efficient, will then be avoided. Considering a frequency sweep in the order of 1 GHz and taking production spread into account, some 8% of fractional bandwidth for the antenna is required. In this view the choice of a printed dipole element can be easily justified. It does however imply the need for a backing reflector to dramatically increase the front-to-back ratio. The supplied differential lines have a characteristic impedance of 100 Ω. The lines are connected to the differential input and output of the QFN package. As the RF input and output are located at adjacent sides of the package, see Fig. 4, a bend in each feeding line of the equally polarized antennas is mandatory.

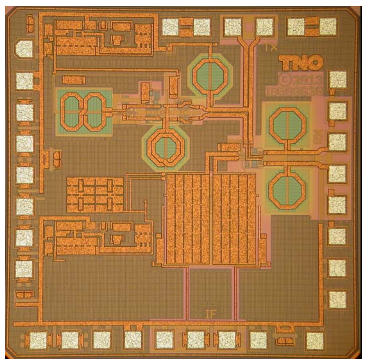

Fig. 3. A microphotograph of the MMIC transceiver (1.1x0.9 mm^2).

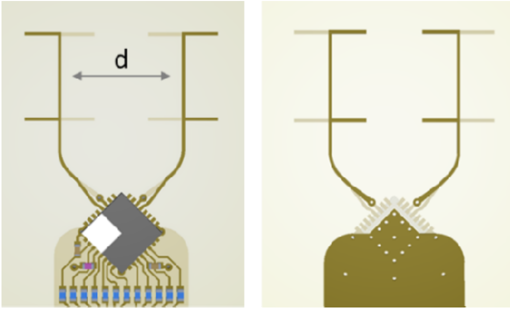

Fig. 4. MMIC, circuitry and antenna array pair: top side view (left) and bottom side (right) also showing the ground for the electronics.

Where the average beam width of 50° can in one plane (E-plane) be achieved using a single dipole element, it does require additional measures in the orthogonal plane. Related to this is the choice of transmission line to excite the antennas, which is a trade-off between beam width and matching. Utilizing a coplanar strip line (CPS) to feed the antenna elements is a logical step to consider. But to obtain the wanted impedance for such a line, the necessary gap and track widths approach the typical etching process tolerances. Additionally, the mandatory (multiple) half wavelength separation between the dipoles does not create the wanted beam width. If a parallel strip line (PS) is chosen instead, the field lines are trapped mainly in the dielectric. This gives additional control of the in-channel dipole separation through the substrate permittivity and thickness. With the choice of RO4350 (10 mil) as the substrate material, the H-plane beam width can be obtained with sufficient matching. Applying then two dipoles per channel suffices. Only a very limited portion of the substrate is devoted to efficiently transform from CPS to PS. A schematic view of the resulting antenna arrays is presented in Fig. 4.

For proper functionality of the sensor a cross talk level between the channels lower than -25 dB is necessary. For the current setup this level is mainly influenced by the distance (*d*, Fig. 4) between the antenna arrays. In Fig. 5 the calculated and measured reflection and cross talk levels are indicated.

The antennas require the presence of a backing reflector at about 3 mm distance behind the arrays. The thus induced radiation is perpendicular to the substrate. Separation *d* between the arrays is 11 mm. For the sake of clarity only the calculated radiation patterns in the principal planes are presented at the center frequency of 24 GHz, refer to Fig. 6. It is mentioned however, that over the frequency range 23 – 25 GHz, the patterns show a well formed and stable main beam. A single array realizes at these frequencies an average calculated gain of 10 dBi.

V. Measurements

As the characterization of the antenna configuration has already been treated in the previous section, this section focuses on the MMIC. The packaged MMIC is characterized on a connectorized PCB test board, see Fig. 7. Decoupling capacitors of 100 nF are placed close to the device on the VCC_{tx} and VCC_{rx} power-supplies. The V_{tune} port is decoupled with a 10 pF capacitor.

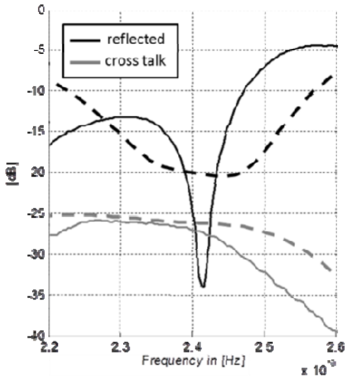

Fig. 5. Calculated (dashed) and measured (solid) reflected and cross coupled power.

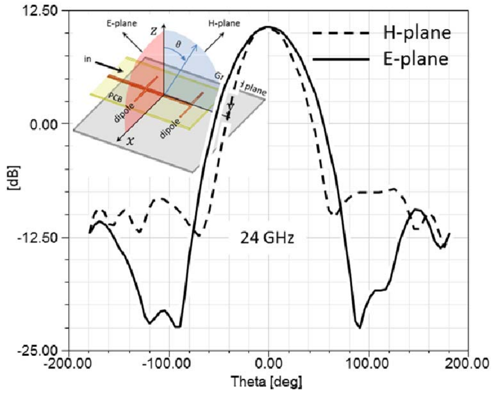

Fig. 6. Calculated radiation patterns (realized gain) at the center frequency in both principal planes. Main beam oriented along the z-axis.

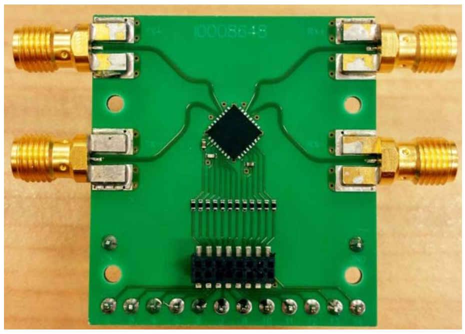

Fig. 7. PCB test board for the packaged MMIC characterization.

The TX part has been measured, with one of the output ports connected to the measurement system, all un-used RF ports are terminated with 50 Ω loads. The oscillation frequency, phase-noise and pulling are measured with a spectrum analyzer. The output power is measured using a power meter. The output power is -13 dBm for a single port and is compared with the simulations in Fig. 8. The output power is well predicted by the simulations. The oscillation frequency as a function of tune voltage has been measured for two settings of the frequency band. This frequency selection was adopted to compensate for a frequency shift due to process variation. The measured frequency is compared with the simulation in fig 9. It shows that the measured frequency has been shifted down by 1 GHz. The TX parameters have been measured as a function of temperature, by placing the test board inside a temperature controlled chamber. The oscillation frequency decreases 6 MHz/°C, as is depicted in fig 10. This is well predicted by the simulations, which show a decrease of 5 MHz/°C.

To measure the RX performance of the MMIC, the on-chip VCO has been used as the LO. Due to the free-running nature of the VCO, it is very difficult to attain a stable beat-note frequency using an external RF source at the receiver input. The integrated phase-noise causes a frequency jitter, which completely covers the beat-note frequency range. Due to the relatively low external Q of the VCO it is possible to stabilize the VCO by injection-locking it to a second external RF source, which is injected via the TX port. To connect the IF output port to the 50 Ω input of the E4446 spectrum analyzer an output network is required, which consist of a blocking capacitor, a series-resistor and a attenuator. The loss of this network is low enough to maintain an overall conversion gain, which is high enough to keep the noise-floor of the spectrum analyzer well below the measured noise-level. Both signal level and noise level (calibrated in dBm/Hz) can directly be measured with the spectrum analyzer. The results are compared with the simulations and depicted in fig 11. The noise figure is 22.9 dB at 6 kHz and the maximum gain is 51.5 dB with a corner frequency of 6.3 kHz of the HPF. Although other work [4-6] is highly integrated, this chip has a very low power consumption of 86 mW and an integrated IF amplifier for direct processing with an ADC.

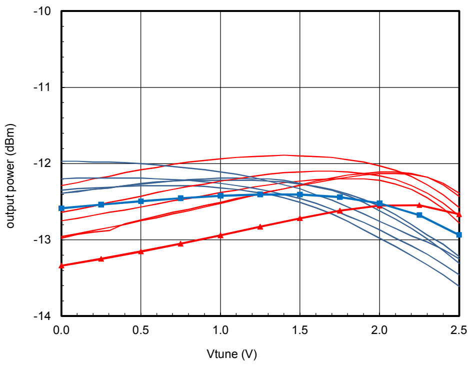

Fig. 8. Comparison between measured and simulated TX output power for 0V (blue, square) and 3.3V (red, triangle) of the MOSCAP source voltage. Marked traces represent the simulation results.

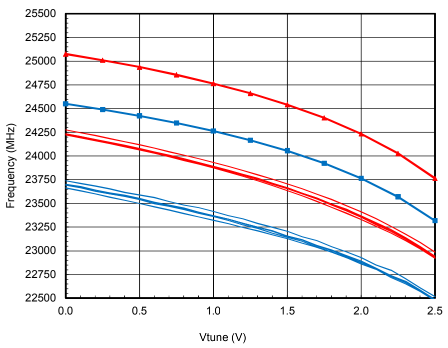

Fig. 9. Comparison between measured and simulated oscillation frequency versus varicap tune voltage for 0V (blue, square) and 3.3V (red, triangle) of the MOSCAP source voltage. Marked traces represent the simulation results.

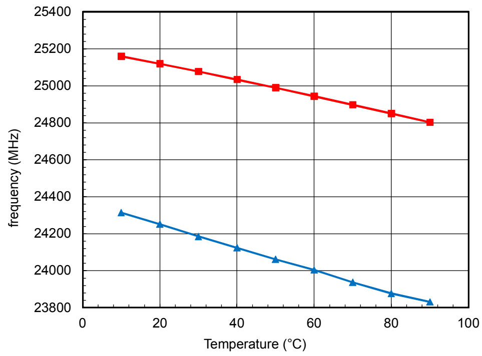

Fig. 10. Comparison of simulated (square) and measured (triangle) oscillation frequency as a function of temperature.

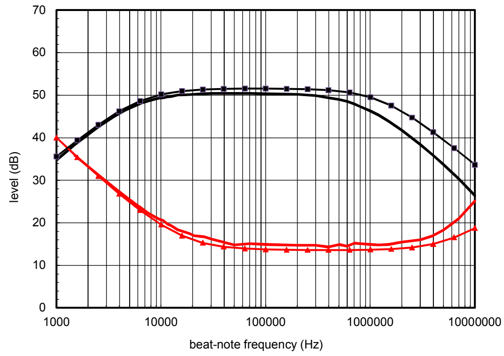

Fig. 11. Comparison between measured and simulated RX noise figure (red, triangle) and conversion gain (black, square). Marked traces represent the simulation results.

VI. Conclusion

In this paper a first time right 24 GHz FMCW radar transceiver is presented. The complete radar is low cost and consist only of three components: a commercial ARM microcontroller, a MMIC front-end and a PCB dipole antenna array. The low power front-end consumes 86 mW from a 3.3 V power supply. On account of the front-end performance of -10 dBm output power; 10 dBi antenna gain; 51 dB conversion gain and 22 dB noise figure, the signal processing can be implemented on an ARM C3 processor. This results in a small presence detection sensor required to realize intelligent LED lighting systems.

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