Low-Cost GaAs pHEMT MMIC's for Millimeter-Wave Sensor Applications

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***Abstract*—A set of coplanar monolithic microwave integrated circuits for millimeter-wave sensor applications is described. It consists of a highly integrated transceiver chip, a voltage-controlled oscillator, a harmonic mixer, and a general-purpose medium-power amplifier. The circuits operate in the 76–77 GHz frequency range and have been fabricated by a production-oriented GaAs pHEMT technology. The transceiver chip combines a transmitter with 9-dBm output power and a receiver with an overall conversion gain of 1 dB. The voltage-controlled oscillator is tunable over a 0.8 GHz bandwidth. It includes a buffer amplifier and generates an output power of 10 dBm. The harmonic mixer achieves 18 dB conversion loss when mixing with the fifth harmonic of the LO signal. The two-stage MPA delivers 13 dBm of output power along with a gain of 7.5 dB. The chip set is suited for the cost effective realization of automotive radar systems as well as various sensors for industrial applications.**

***Index Terms*—Millimeter-wave FET integrated circuits, mixer noise, MMIC amplifiers, MMIC mixers, MMIC oscillators, MODFET integrated circuits, road vehicle radar.**

I. Introduction

SENSOR systems are among the most important commercial applications of millimeter-waves [1]. In particular, the forward-looking automotive radar for purposes of collision warning and autonomous intelligent cruise control (AICC) is expected to emerge into a high-volume market within the next few years. A promising way to meet the stringent cost requirements of these systems is the use of monolithic microwave integrated circuits (MMIC's) based on gallium arsenide (GaAs) pseudomorphic high electron mobility transistor (pHEMT) technologies. A complete W-band FWCW radar front end has already been integrated on a single pHEMT chip with an area of less than 25 mm^2 [2]. This MMIC was employed in a prototype single-beam automotive radar with a range of about 100 m and a performance adequate for AICC [3]. However, for the design of multiple-beam systems, a set of chips offers more flexibility than a single-chip approach. A chip set covering all functions of a homodyne FMCW radar front end operating in the 76–77 GHz frequency range has been presented in [4]. It consists of a voltage-controlled oscillator (VCO), an active harmonic mixer, and two highly integrated circuits comprising the essential functions of the transmitter and receiver path, respectively. The overall cost is minimized by compact coplanar chip designs and a production-oriented pHEMT technology.

This paper describes significant improvements of that chip set. The main difference is a new transceiver chip, which integrates most of the functions of the formerly separate transmitter and receiver MMIC's. It is supplemented by a new medium-power amplifier (MPA), which may be used as a general-purpose gain block, and redesigned versions of both the voltage-controlled oscillator and the harmonic mixer. Compared to the first-generation designs [4], these new GaAs MMIC's make sensor systems possible with less total chip area and thus lower cost. Moreover, they are particularly suited for system architectures with a common transmit and receive antenna. Single antenna systems best comply with the demand for low dimensions of the sensor module, a critical aspect in automotive applications. The chip set may also be applied to other commercial sensor systems, as, for instance, in the fields of motion detection or industrial level measurements.

II. Fabrication Technology

The MMIC's have been fabricated using the Siemens HEMT110 process technology. The number indicates that the active devices show a current-gain cutoff frequency of 110 GHz at normal dc operating conditions, i.e., at a drain-to-source voltage of 2 V and a normalized drain current of 250 mA/mm. From measured S-parameters, a maximum frequency of oscillation above 200 GHz has been extrapolated. The maximum extrinsic transconductance is beyond 700 mS/mm, the gate-to-drain breakdown voltage is better than 5 V, and the maximum normalized drain current under open channel conditions exceeds 600 mA/mm. From these numbers, an output power capability of 150 mW/mm is estimated at millimeter-wave frequencies. Minimum noise figures of 0.6 and 1.5 dB have been measured at 12 and 26 GHz, respectively.

The HEMT110 process is an enhanced version of a well established and space qualified fabrication technology for lower frequency applications [5], [6]. The delta-doped double-heterojunction pseudomorphic active device layers are grown in-house by molecular beam epitaxy on 3-in GaAs substrates. Optical stepper lithography is applied throughout the whole

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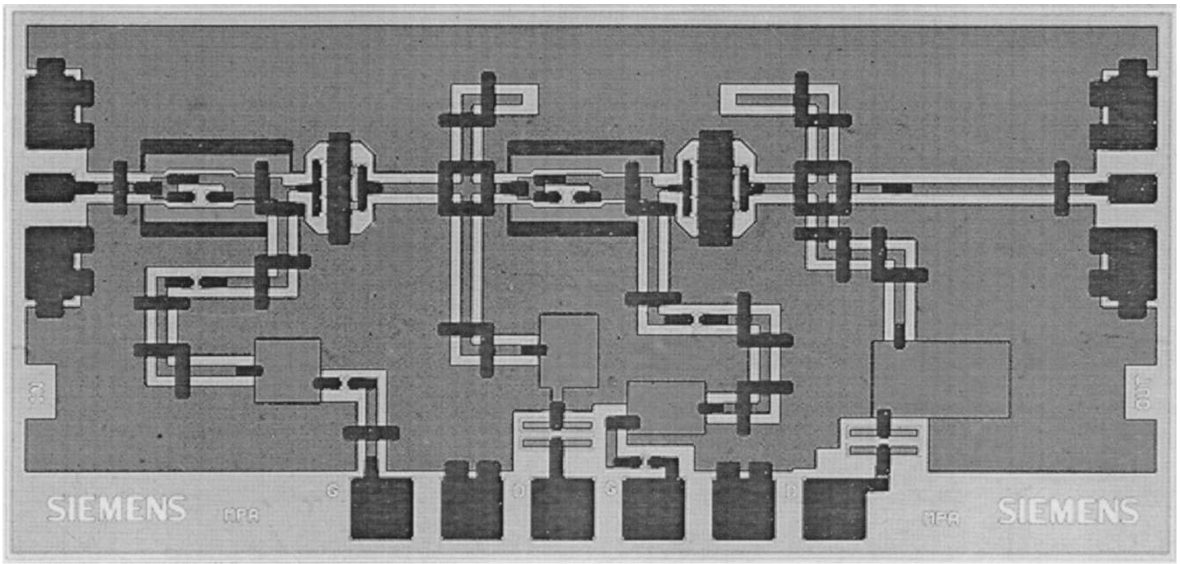


Fig. 1. Chip photograph of the two-stage MPA (1.0 mm × 2.0 mm).

process including the critical gate formation. A gate length of 0.12 μm with typically ±5% variation across a wafer is achieved by using phase-shift masks and a sidewall spacer process. The T-shaped gate consists of a refractory metal and a highly conductive gold overlay, which yields high reliability and low gate resistance. Ohmic resistors are made either by utilizing the cap layer of the epitaxial structure or, as a high precision option, by an additional NiCr thin film. The MMIC process is completed by metal–insulator–metal (MIM) capacitors and a total of three interconnect metal layers including electroplated airbridges. The process is described in more detail in [4].

Since this technology, in contrast to most other fabrication methods for high performance devices, does not rely on electron beam lithography, it is suited for the high-throughput production of low-cost millimeter-wave integrated circuits. As a further step toward higher productivity, the chip fabrication is currently transferred to equipment capable of processing 4-in wafers.

III. Chip Design and Measurement Results

All circuit designs use coplanar waveguide transmission lines in order to achieve compact dimensions and to avoid the need for backside metallization and via holes. The modeling of the coplanar elements was based on the results described in [7]. An improved nonlinear HEMT model [8] has been implemented into a commercial CAD system for large-signal simulations.

A. MPA

A photograph of the newly developed two-stage MPA is shown in Fig. 1. The size of the chip is 1.0 mm × 2.0 mm. Each of the two transistors has four gate fingers. The total gate width is 160 μm for the first-stage HEMT and 240 μm for the second-stage device. The matching networks consist of coplanar transmission lines including open stubs. Furthermore,

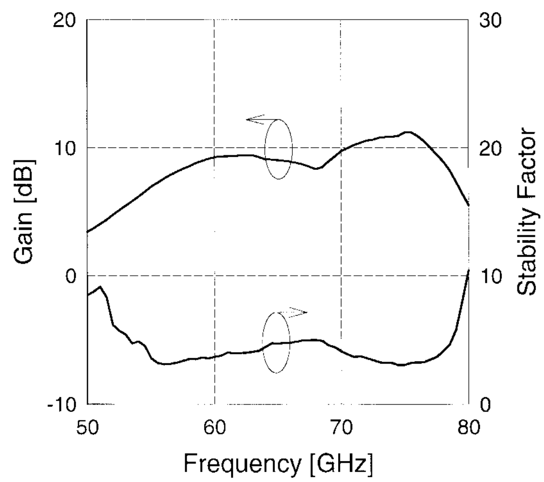


Fig. 2. Measured small-signal gain and stability factor of the MPA.

resistor-capacitor combinations are inserted in the gate transmission lines of both transistors. The attenuation introduced by these networks below the nominal operating frequency helps to avoid out-of-band oscillations and to enhance the overall stability of the amplifier. Low-frequency oscillations are suppressed by low-pass filters in each of the four bias voltage lines. Both stages are operated at 3 V drain-to-source voltage with drain dc currents of 40 and 60 mA, respectively.

Measured curves of the small-signal gain and of the stability factor over the frequency range 50–80 GHz are depicted in Fig. 2. The maximum gain is about 11 dB at 76 GHz. The corresponding return loss is better than 10 dB at both ports. The amplifier gain shows, however, a broadband performance. Thus, potential applications are not restricted to the 76–77 GHz frequency band. In particular, the amplifier still offers useful gain at 60 GHz, another important frequency for future commercial millimeter-wave systems. The emphasis placed on stability considerations during the circuit design is evident from the large values of the stability factor. The amplifier is unconditionally stable at all frequencies.

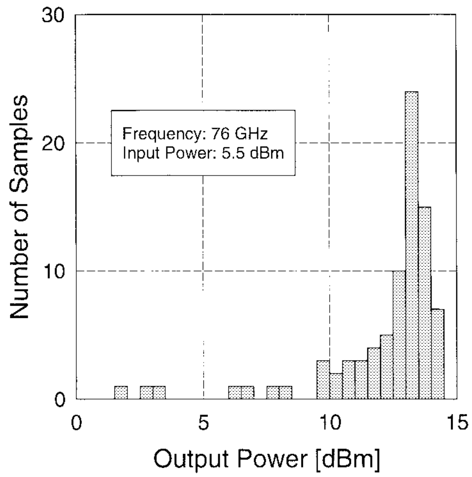


Fig. 3. Measured output power distribution of the MPA chips from one wafer.

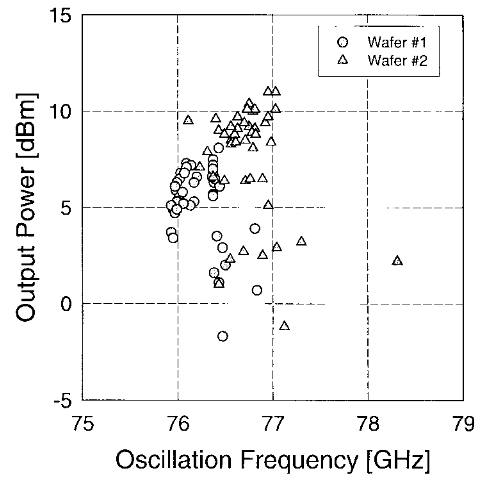


Fig. 5. Measured output power and oscillation frequency of the oscillator chips from two wafers of different lots.

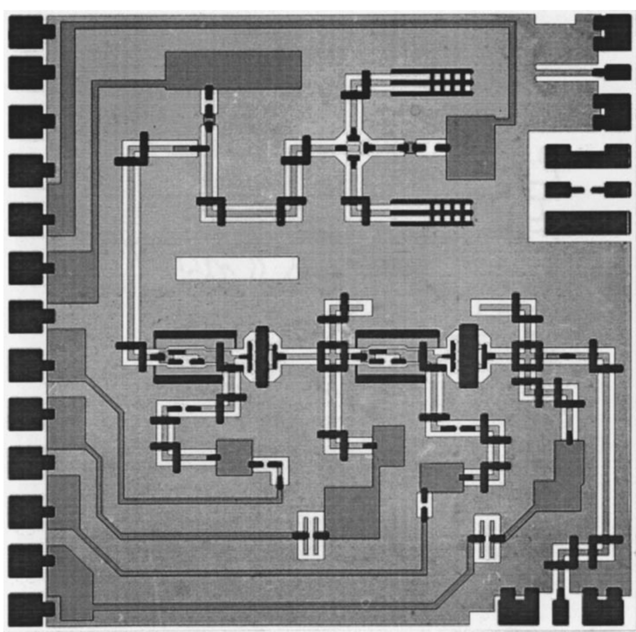


Fig. 4. Chip photograph of the VCO with buffer amplifier (2.0 mm × 2.0 mm).

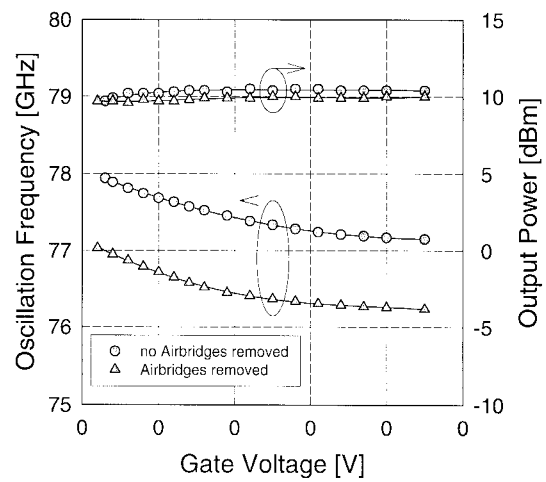


Fig. 6. Measured tuning characteristics of the voltage-controlled oscillator.

The diagram in Fig. 3 summarizes the results of an automatic large-signal characterization of the amplifier across a complete 3-in wafer. At 76 GHz, the output power has been measured with a fixed input power level of 5.5 dBm. The peak of the distribution is located at an output power of 13 dBm corresponding to a power gain of 7.5 dB and a gain compression of 3.5 dB. Taking into account a total number of 95 chips on the wafer and a specified minimum output power of 12 dBm, a yield of 64% is obtained from the data in Fig. 3.

B. Voltage-Controlled Oscillator

The redesigned version of the VCO is shown in Fig. 4. The oscillator stage is located in the upper half of the picture, the remaining part of the chip area is assigned to the buffer amplifier. In addition to some design modifications of the oscillator circuit itself, the former one-stage buffer amplifier has been replaced by the new two-stage MPA. This results in higher output power and less susceptibility to load impedance variations. The dimensions of the chip are 2.0 mm × 2.0 mm.

The active device in the oscillator stage is a two-finger HEMT with 80 μm gate width. Gate, drain, and the two source contacts of the transistor are connected to coplanar transmission lines, all of which are shorted to ground at the opposite end, either by airbridges or by MIM capacitors. The lengths of these lines determine both the loop gain and the oscillation frequency. In case of the two source lines, the length may be changed by altering the position of the shorting airbridges. This offers a simple method to compensate for deviations of the oscillation frequency from the specified value without major modifications of the layout. One way to change the short-circuit position is by using a modified mask for the airbridge lithography, which is one of the final processing steps. This might be appropriate to compensate for HEMT parameter variations, which can be measured at special test transistors after deposition of the ohmic metal during the fabrication process. Increasing the effective line length is even possible after the wafer process has been completed. As can be seen in the chip photograph, several airbridges are

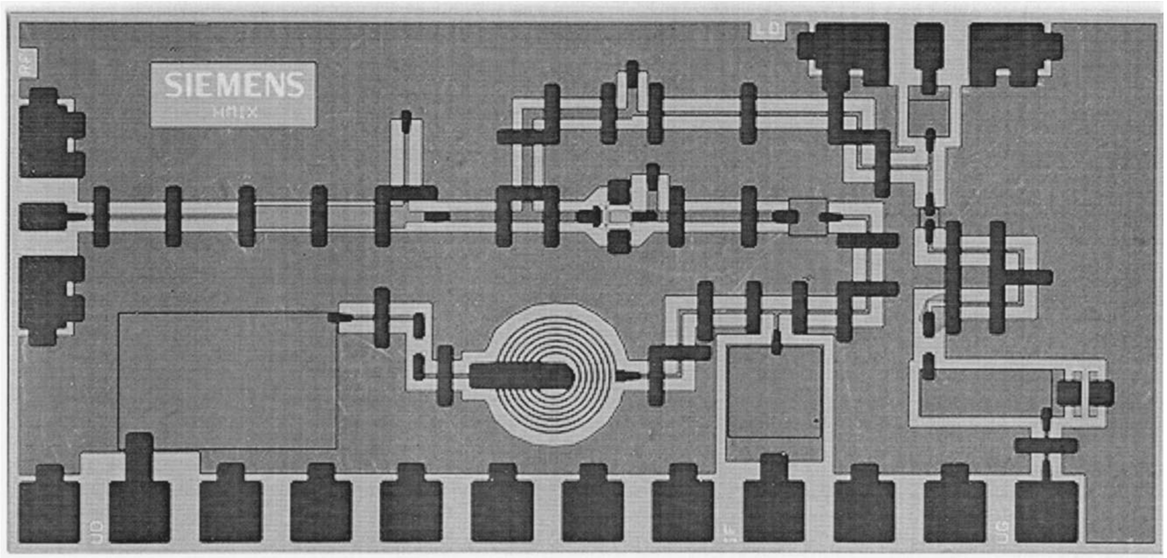


Fig. 7. Chip photograph of the active harmonic mixer (1.0 mm × 2.0 mm).

successively placed at the end of the source lines. The lines are shorted by the airbridges located next to the HEMT source electrodes. If the first bridges are mechanically removed, the next airbridges take over the short-circuit function, which results in an increased electrical length of the lines. In this way, a step-by-step tuning toward lower oscillation frequencies is possible. Electronic tuning of the oscillator is accomplished by varying the gate dc voltage of the active device. A separate tuning varactor is not used. The drain transmission line is tapped close to the shorted end in order to extract the output signal, which then is routed to the input port of the buffer amplifier.

Fig. 5 shows test results for two wafers of different lots. The HEMT in the oscillator stage was operated at a drain-to-source voltage of 3 V with a typical drain dc current of 12 mA. Out of the 95 available chips per wafer, 46 were oscillating on the first and 44 on the second one. In the diagram, the measured output powers are plotted against the respective oscillation frequencies. For the majority of oscillating circuits from wafer #1, output powers in the range of 5–8 dBm and frequencies between 75.9 and 76.5 GHz have been measured. The chips from wafer #2 delivered higher output power, typically 7–11 dBm, but the range of oscillation frequencies is shifted up to values between 76.3 and 77.1 GHz for most of the samples.

The frequencies shown in Fig. 5 were measured at the lower edge of the electronic tuning range. By lowering the gate dc voltage of the oscillator transistor, the frequency can be raised by typically 0.8 GHz. Fig. 6 shows the electronic tuning characteristics of an oscillator chip with one of the highest oscillation frequencies from wafer #2. The circles denote the measurement points for the circuit in its original state. The entire tuning curve is located above 77 GHz, i.e., out of the specified frequency band. In the way described above, the oscillator was subsequently trimmed by removing airbridges at the source transmission lines. The measurement results obtained after the end of this procedure are marked by the triangles in Fig. 6. The electronic tuning range now extends

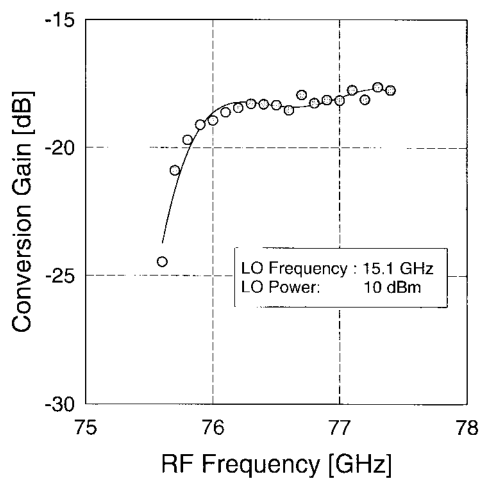


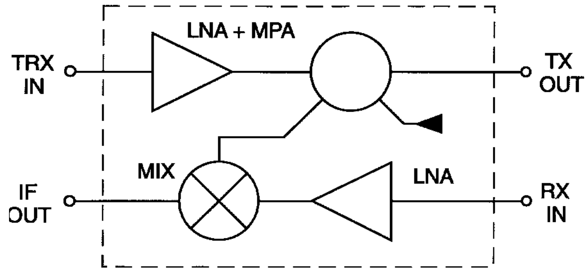
Fig. 8. Measured conversion gain of the harmonic mixer.

from 76.2 to 77.0 GHz. The correction of the oscillation frequency only slightly affects the output power. In both cases, about 10 dBm are available at the output of the buffer amplifier. If performed in an automatic fashion by means of a laser, this kind of trimming procedure might be applicable in a production environment to compensate for technology-induced frequency deviations and thus to enhance the yield of the VCO.

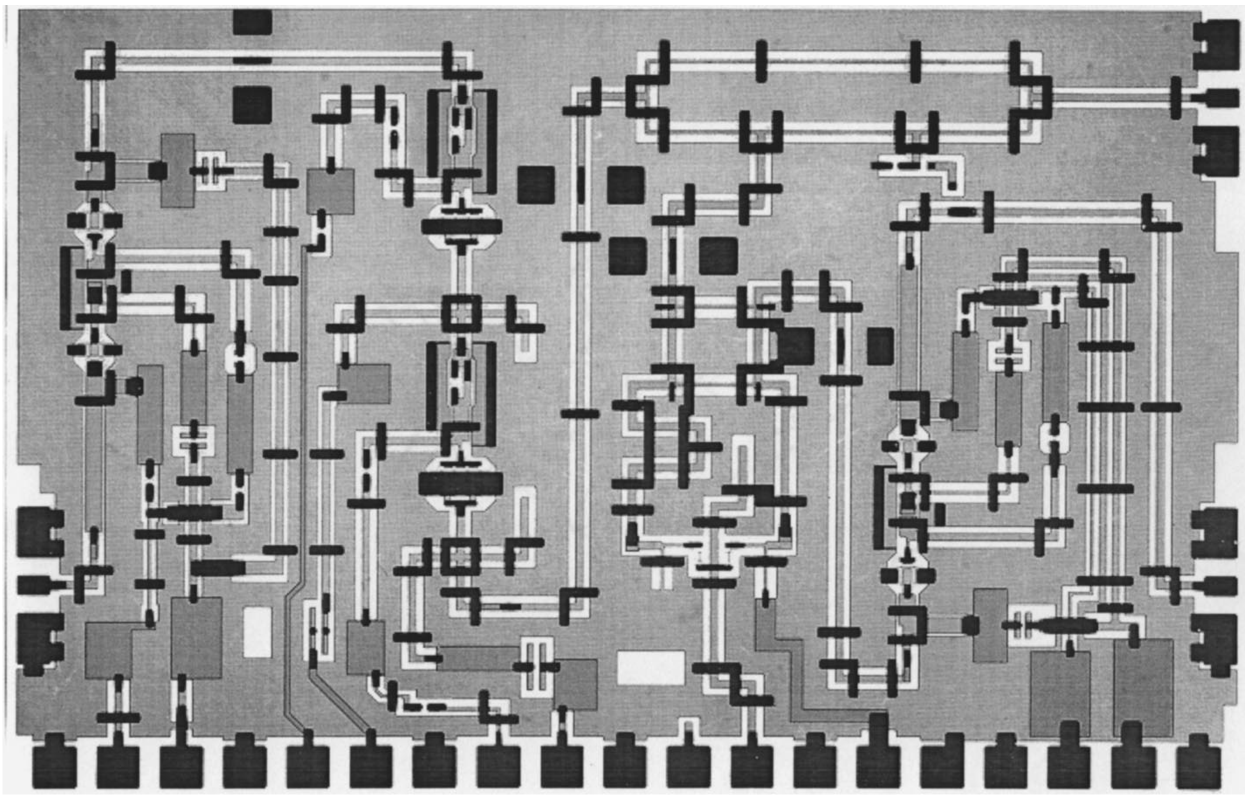
The typical phase noise of the VCO, as estimated from spectrum analyzer measurements, is −80 dBc/Hz at 1 MHz offset frequency.

C. Harmonic Mixer

The active harmonic mixer (HMIX), shown in Fig. 7, has a size of 1.0 mm × 2.0 mm. A two-finger HEMT with 80-μm gate width is used as the nonlinear active device. It is operated at a drain-to-source voltage of 3 V. Both RF and local oscillator (LO) signals are applied to the gate via coplanar decoupling and matching networks. The low-



(a)



(b)

Fig. 9. (a) Block diagram and (b) chip photograph of the transceiver MMIC (2.0 mm × 3.0 mm).

frequency IF signal is extracted from the drain circuit. The IF path is isolated from the drain bias line by a filter network including a circular spiral inductor. The inductance of this component largely determines the lower edge of the useful IF frequency range. Due to insufficient accuracy of the spiral inductor model used in the first design, the mixer performance suffered from an increased conversion loss at IF frequencies below 1 GHz. This problem has been fixed with the redesign. Fig. 8 shows the measured conversion gain as a function of the RF frequency using a fixed LO signal at 15.1 GHz with a power level of 10 dBm. When mixing with the fifth harmonic of the LO signal, the RF frequency range of 76–77 GHz is converted to the IF frequency band 0.5–1.5 GHz. An almost constant conversion loss of 18–19 dB has been measured for this mode of operation. The conversion loss exceeds 20 dB only if the IF frequency is lower than approximately 300 MHz. A detailed description of the mixer design including additional measurement and simulation results is given in [9].

D. Transceiver Circuit

A block diagram and a photograph of the transceiver (TRX) which has dimensions of 2.0 mm × 3.0 mm are shown in Fig. 9. The receiver input port is connected to a newly designed two-stage low-noise amplifier (LNA), located in the lower right part of the chip. The gate widths of the two HEMT's are 60 and 80 μm, respectively. The amplifier design was optimized for high gain in the 76–77 GHz band. Thus, the resulting bandwidth is lower than in the case of the MPA. The preamplifier is followed by a single-balanced mixer, comprising a 3-dB branch line coupler and two Schottky diodes. The diodes actually are HEMT's with a single 20-μm wide gate finger and combined drain-source contacts. A dc bias current can be applied to the diodes in order to reduce the LO power requirements. In the chip photograph, the mixer is located next to the preamplifier. The entire left half of the chip consists of a four-stage transmit amplifier, which is a cascade connection of the LNA and MPA circuits. Via a rat-race coupler, which can

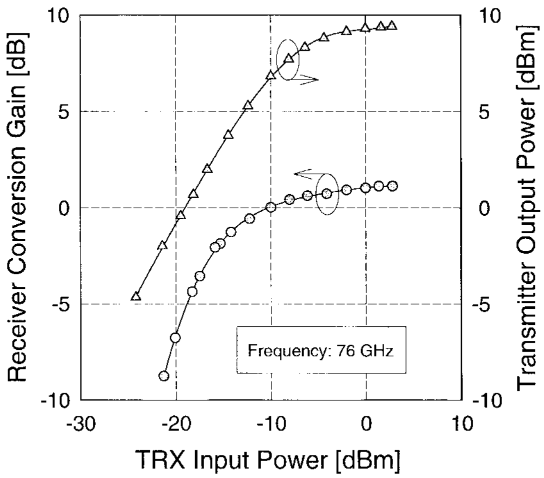


Fig. 10. Measured receiver conversion gain and transmitter output power of the transceiver.

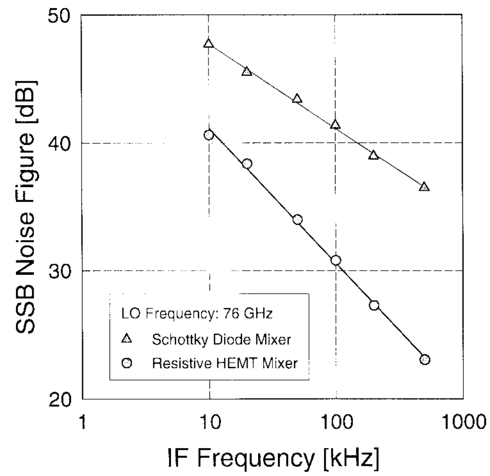


Fig. 11. Measured single-sideband noise figures of balanced diode and resistive HEMT mixer circuits as a function of IF frequency.

be seen as a rectangle in the upper right corner of the chip, half of the amplifier output power is delivered to the transmitter output port, the other half serves as the LO signal for the mixer. The fourth port of the coupler, normally terminated by a 50-Ω resistor, can alternatively be connected to the preamplifier input port by minor airbridge modifications. In this case, the transmitter output port is also utilized as the receiver input port with the coupler taking over the additional function of a transmit-receive diplexer. By this option, the transceiver can easily be adapted to different antenna feed configurations.

Measured results of the transceiver are shown in Fig. 10. For a fixed frequency of 76 GHz, the transmitter output power and the receiver conversion gain are plotted versus the input power of the four-stage amplifier. The conversion gain saturates at around 1 dB, indicating that the LNA gain is slightly higher than the loss of the mixer. With a -3-dBm input signal, 9 dBm of output power is available at the transmitter output port. This is in accordance with the measurement results of the MPA (see Fig. 3), taking into account the 3-dB ideal coupling factor of the rat-race coupler plus an additional loss of 1 dB. The results of corresponding measurements at 77 GHz were almost identical to the data shown in Fig. 10.

For two reasons the mixer is a key component of the transceiver circuit. Due to the homodyne system architecture, the IF signals are at very low frequencies, typically below 1 MHz for an FMCW radar. Therefore, the receiver sensitivity and thus the sensor range largely depend on the amount of $1/f$-noise generated in the mixer. The receiver sensitivity can be improved by using an LNA in front of the mixer, as shown in the block diagram of Fig. 9. However, the LNA gain may not exceed a certain value in order to avoid nonlinear signal distortions in the mixer. Since an automotive radar, for instance, must be capable of handling multiple-target situations, a large intermodulation-free dynamic range of the receiver is mandatory. This demand may become even more severe due to the limited isolation between the transmitter and receiver paths. In summary, both the $1/f$-noise and the linearity of the mixer to a great extent determine the attainable sensor performance.

In addition to the conventional Schottky diode mixer, several other coplanar monolithic mixer circuits have been investigated with respect to millimeter-wave sensor applications [10]. Promising results were found for resistive HEMT mixers. The basic circuit concept is known primarily because of its excellent intermodulation performance [11]. Moreover, since the devices are operated in a passive mode without a drain dc current, less $1/f$-noise should be generated than in other mixer circuits. This is confirmed by Fig. 11, which shows measured single-sideband (SSB) noise figures as a function of the IF frequency. The two curves correspond to different mixer circuits, namely a diode mixer like the one implemented in the transceiver chip of Fig. 9 and a balanced resistive mixer with two HEMT's of 30-μm gate width each. Both mixers have been fabricated on the same wafer. While nearly the same conversion loss of approximately 11 dB was measured for the two circuits, the noise figures of the resistive mixer are significantly lower. Taking a noise difference of 11 dB at 100 kHz IF frequency, the range of a radar sensor can almost be doubled by using the resistive mixer instead of the Schottky diode type.

Because of these results, a second version of the transceiver chip has been designed with a resistive HEMT mixer replacing the diode mixer [12]. The other parts of the MMIC as well as the chip dimensions remained unchanged. Accordingly, the measured values of transmitter output power and receiver conversion gain were very similar to the corresponding data of the first transceiver version shown in Fig. 10.

IV. Applications

Automotive radar systems are regarded as the most important field of application for this chip set. As an example, Fig. 12 shows the block diagram of a possible FMCW radar front end. One transceiver is assigned to each of three transmit-receive beams. The three beams point to slightly different azimuthal directions. This enables the determination of the angular position of detected objects by evaluating the three different IF signals. Separate transmit and receive antenna feeds are assumed here which, however, share a single dielectric lens to form a compact high gain antenna. The input signals of the transceivers are supplied by the VCO via a

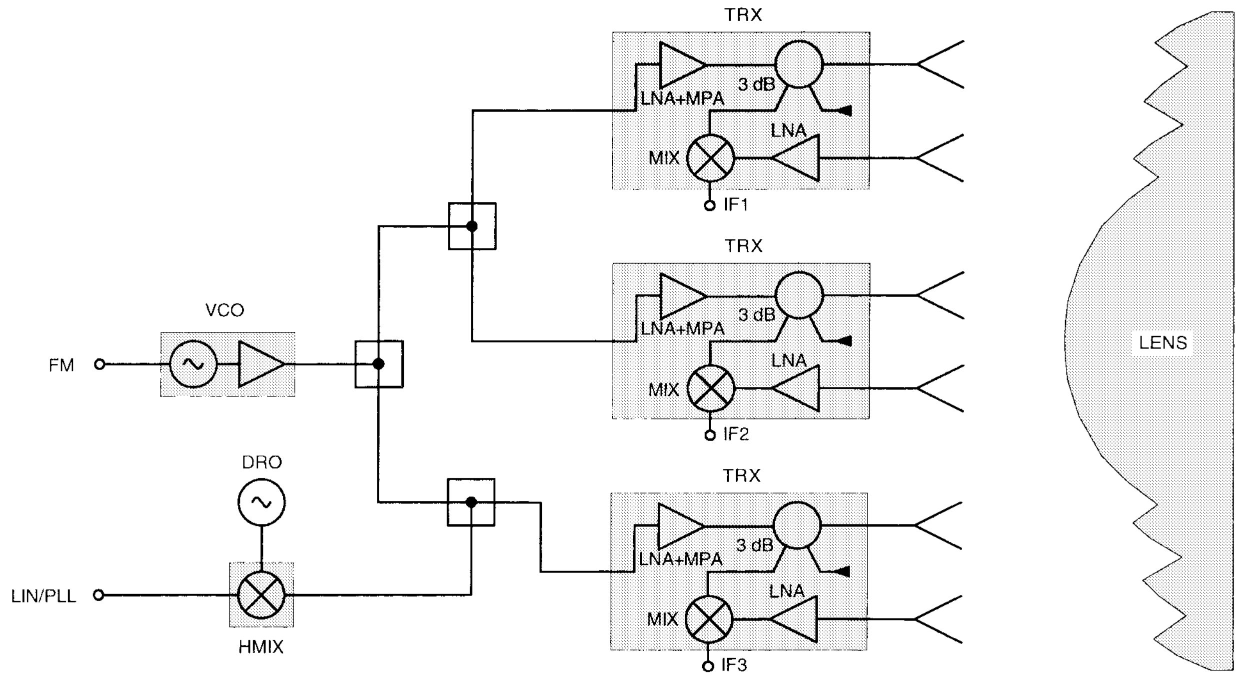


Fig. 12. Block diagram of a three-beam FMCW radar front end as an application example of the chip set.

passive distribution network. A part of the oscillator signal is downconverted to a low frequency by the harmonic mixer and a stable LO signal, generated by a dielectric resonator oscillator (DRO). The downconverted signal can be further processed for VCO linearization and stabilization purposes in order to meet the stringent signal source requirements of a high-resolution FMCW radar. The total GaAs chip area required for this system is only 24 mm^2. Based on the measured results of the transceiver chip with the resistive mixer, the performance of this radar sensor has been estimated. If an antenna gain of 30 dB, a receiver noise figure of 22 dB at 100 kHz IF frequency, and an IF resolution of 0.5 kHz are assumed, then a target with 1 m^2 radar cross section at a distance of 150 m should be detected with a signal-to-noise ratio of better than 15 dB. This appears to be adequate for a forward-looking automotive radar.

The block diagram in Fig. 12 can be modified in a variety of ways. A better angular resolution of the radar can be obtained if the number of antenna beams and accordingly the number of transceiver chips is increased. Losses of the distribution network may be compensated by inserting samples of the MPA. If, on the other hand, a single-beam system already meets the sensor requirements, the number of MMIC's can be reduced to three with a total chip area of 12 mm^2. A simple short-range motion detector or speed sensor may be realized with only two chips just by combining the VCO chip with one transceiver circuit, since the oscillator linearization loop including the DRO and the harmonic mixer can be omitted in a pure Doppler radar.

V. Conclusion

Several millimeter-wave MMIC's have been developed for sensor applications in the 76–77 GHz frequency range. The chip set includes a VCO generating a 10-dBm signal tunable over a 0.8-GHz bandwidth, a harmonic mixer with 18 dB conversion loss, and a MPA with 13 dBm of output power. The most complex circuit is a transceiver chip which achieves 9 dBm of transmitter output power and 1 dB receiver conversion gain. Two versions of the transceiver have been designed, one with a Schottky diode mixer and a second one with a resistive HEMT mixer. The latter version is expected to have an increased receiver dynamic range corresponding to better sensor performance. Coplanar waveguide transmission lines are employed in all circuit designs, resulting in compact chip dimensions.

The monolithic circuits have been fabricated by an advanced GaAs pHEMT technology featuring active devices with 0.12-μm gate length and 110-GHz cutoff frequency. The process is suited for high-volume production since no electron beam lithography is required. The statistical evaluation of the available measurement data indicates a potentially high yield of the MMIC fabrication. In case of the VCO, the yield can be further enhanced by mechanical trimming of the circuit to coarse adjust the oscillation frequency.

In summary, the high-yield production-oriented process technology in conjunction with the compact designs are expected to result in low cost of the monolithic circuits when manufactured in large quantities. The chip set thus could become a basis for the cost effective realization of various millimeter-wave sensor systems for automotive and industrial applications.

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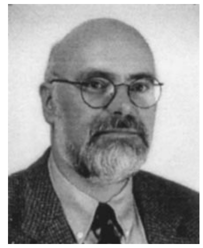
From 1979 to 1985, he was with the Institut für Hoch- und Höchstfrequenztechnik of the Ruhr-Universität, where he worked on radiometer systems, frequency multipliers, and GaAs MESFET oscillators. In 1985, he joined Siemens Corporate Research and Development in Munich, Germany. He has been involved in the characterization and modeling of GaAs devices and was responsible for the design of various MMIC components based on MESFET and HEMT technologies. Currently, he is engaged in the development of automotive radar systems employing millimeter-wave HEMT MMIC's.

Andreas Werthof was born in Kassel, Germany, in 1961. He received the Dipl.-Ing. and Dr.-Ing. degree from the University of Kassel in 1988 and 1994, respectively.

In 1988, he joined the Siemens Defence Electronics Group, where he was involved in the development of a transmit/receive module for an X-band phased array radar. In 1989, he returned to the University of Kassel to work on his thesis on large signal circuit models for FET's and HBT's. In 1994, he joined the Ferdinand–Braun Institute in Berlin, where he developed a large signal model for Si/SiGe HBT's. Since June 1995, he has been with Siemens Corporate Technology working on HEMT modeling and on HEMT circuit design for radio links and automotive radar.



Harald Tischer was born in Delmenhorst, Germany, in 1952. He received the diploma degree in physics and the Dr.rer.nat. degree from the University of Göttingen in 1978 and 1982, respectively.

During his thesis work, he studied the scattering and interaction potentials of electronically excited alkali with rare gas atoms. From 1983 to 1985, he worked as a Postdoctoral Research Associate at JILA, Boulder, CO, on the collisional quelching of metastable hydrogen molecules. He joined the Siemens Corporate Technology Department in 1986, where he first worked in the field of optical communications on Gb/s transmission systems and the characterization and modeling of laser diodes. In 1993, he joined the III/V electronics group and is currently engaged in the characterization of GaAs HEMT's and MMIC's up to millimeter wave frequencies.

Ulrich Schaper (M'91) received the diploma in physics from the University of Munich in 1976 and the Ph.D. degree from the Technical University of Munich in 1981.

He worked at the Max–Planck Institute for Plasma Physics in Garching near Munich on nuclear fusion research. He spent the following years as an Assistant Researcher at the Institute for Theoretical Physics of the University of Düsseldorf, concerned with the stability of plasma equilibria for nuclear fusion applications. In 1984, he joined Siemens Components Group in Munich working on SPICE models for MOSFET's. Since 1986, he has been with Siemens Corporate Technology, where he developed various models for HBT's and HEMT's. He also designed mixer circuits for automotive radar applications. Since November 1997, he has been involved in CMOS modeling and circuit design.

A. Schäfer, photograph and biography not available at the time of publication.

L. Verweyen, photograph and biography not available at the time of publication.

T. Grave, photograph and biography not available at the time of publication.

G. Böck, photograph and biography not available at the time of publication.

M. Schlechtweg, photograph and biography not available at the time of publication.

W. Kellner, photograph and biography not available at the time of publication.