High Precision Interferometric Radar for Sheet Thickness Monitoring

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Abstract—Contact-less sensing plays an important role in today's highly automated industrial environment, as modern sensors allow for an increased product quality in high-speed manufacturing lines. Hereby, a measurement system for sheet thickness monitoring in aluminum and steel rolling mills is presented. The presented concept is based on a differential measurement principle using two cooperating six-port radar systems. The radio frequency front-ends are designed in a high density substrate integrated waveguide technology at a frequency of 61 GHz. Furthermore, a detailed analysis of non-ideal behavior of a six-port radar front-ends is presented. Moreover, the measurement principle is evaluated in a simplified measurement setup and the results are compared to previously presented theory. Finally, an overview of modern sensing technology for steel rolling mills is presented and compared to the proposed measurement system considering the important aspects, i.e. update rate, precision, and thickness limit.

Index Terms—radar systems, interferometry, six-port circuits, microwave circuits, substrate integrated waveguide (SIW), planar waveguides

I. INTRODUCTION

Metal processing, e.g., aluminum, iron and steel industry, is a market characterized by intense competition. Therefore, product quality as well as cost reduction have become increasingly important, especially for many European and US American companies. In order to improve yield and quality of produced metal sheets, surface quality as well as their final material thickness, which is also referred as strip exit thickness, are important characteristics to be monitored in rolling mills. Material sheets are typically rolled to their final thickness in several steps. For each operation, the intended sheet thickness is defined by the rolls' gaps. However, roll eccentricity caused by non-uniform thermal expansion or inexact roll grinding can introduce a thickness error of 40 µm and even more in cold and hot rolling mills [1], [2]. Thus, roll eccentricity defines the lower limit of the achievable thickness tolerances. For a further reduction of the tolerances, an active compensation has to be applied to the roll mill [3]. In order to realize such compensation systems, inline sheet thickness sensors are inevitable as roll eccentricity changes with temperature and operating speed [1].

Two different inline sheet thickness measurement systems are well established since more than 50 years: mechanical roll force measurement [4] and X-ray based thickness gauges [3], [5]. Thickness measurement using the roll force method is an indirect measurement method making use of a fix relation between the plastic deformation constant of the processed material, the output sheet thickness deviation and the measured roll force deviation. However, quality related changes of the material parameters influence the metered value [2]. X-ray based gauges are based on measuring the signal power of a nucleonic or X-ray beam transmitted through an absorbing material [5]. Photo detectors are used to obtain the transmitted radiation intensity in form of a voltage value, which is directly related to the material's absorption value as well as the sheet thickness. With today's maximum thickness tolerances being clearly defined with specification values below 0.8 % of the exit thickness, X-ray detectors reach their limit, as they suffer by their nature from a significant random noise component in the detector output signal. Thus, current control technology (roll thickness measurement and X-ray gauges) can not isolate periodic components caused by present roll eccentricity of low amplitude [3].

The described major drawbacks of both systems highlight the need for alternatives. However, the challenge is not only to monitor little changes in thickness and quality of the material, but doing this under difficult conditions, which are the hazardous environment, the high operating speed of typically 20 m/s or higher, and the big varieties of surface defects, e.g. cracks, blowholes, and scratches [6]. Established contact-free sensing technologies like ultrasonic or optical technology can hardly be used in the harsh environment, especially if high update rates are mandatory [7]. Radar systems are perfectly qualified for harsh industrial environment and show competitive performance at reasonable costs [7], [8]. Although classical radar architectures, i.e. pulse radar and frequency modulated continuous wave (FMCW), may feature the recommended accuracy, they are limited in their update rate by nature [9], [10], [11]. In contrast to that, interferometric continuous wave (CW) radar systems (without any modulation) feature high precision displacement detection and extraordinary high update rates at the same time [12], [13]. Therefore, CW radar is a good candidate for monitoring sheet thickness variation in the described application. A special type of CW radar is the six-port reflectometer, using a passive six-port junction with power detectors instead of a conventional
Fig. 1. Sketch of the system design.

mixer [14].

Hereby, a metal sheet thickness monitoring concept based on six-port interferometry is presented and validated in experiment. The publication is organized as follows: Section II will introduce the system concept, followed by a detailed description of the radio frequency frontend design in Section III. Furthermore, this section will give an overview of expectable non-ideal effects in hardware design. The influence of the non-ideal behavior of the six-port radar system on the baseband signals will be analyzed in Section IV. The baseband design and implementation of the sensor for the simplified test setup will be shown in Section V, while Section VI deals with the measurement setup, calibration, and dedicated signal processing. Additionally, the functionality of the sensor will be validated by acquired measurement data. Section VII gives a comprehensive comparison of the presented sensor with state-of-the-art technology. Section VIII finally concludes this paper.

II. SYSTEM DESIGN

A sketch of the system design is displayed in Fig.1. Intending to replace conventional X-ray gauges in existing as well as new rolling mills, the system design basically relies on that of a conventional gauge. Therefore, one or several radar systems pair are mounted opposite on the top and bottom side of the transportation path, where the metal sheet is guided. As a relative measurement of the sheet thickness change is sufficient in the desired application, the equally constructed radar systems are realized as six-port interferometers. Since the frequency $f$ is directly related to the wavelength $\lambda$ by $\lambda = c_0 / f$, with $c_0$ being the speed of light in the surrounding media, a higher system frequency is equivalent to an increased sensitivity regarding small displacements [13]. For millimeter wave frequencies the International Telecommunication Union (ITU) has qualified three license free frequency bands (2.4-2.5 GHz, 24-24.25 GHz, and 61-61.5 GHz) for Industrial, Scientific and Medical (ISM) applications [15]. Furthermore, there are several open frequency bands beyond 100 GHz. In terms of evaluation of the system concept 61 GHz has been chosen as design frequency, as commercial-of-the-shelf components are hardly available at higher frequencies. Thus, in combination with an analog-to-digital converter (ADC), providing an effective number of bits (ENOB) of 18 bits, a resolution in the two digit nanometer range can theoretically be achieved, which is sufficient. The dual antenna concept is necessary by means of the thickness deviation measurement and, furthermore, helps to compensate for vibrations of the moving metal sheet. However, to allow high speed operation, exact positioning of the antenna with an opposite pointing vector is mandatory. Finally, the sampling rate of the 8-channel parallel ADC and the data transmission limits the maximum speed of the sheet movement. For a sheet moving with 20 m/s (787.4 inch/s), assuming a sampling rate of 10 kSa/s, a sample value is obtained every 2 mm (78.74 mil). Although antennas with limited directivity are used, this is much smaller than the illuminated spot size. For this reason, keeping in mind that an integral value is obtained, it is assured that occurring thickness changes will be monitored.

III. RF FRONTEND DESIGN

Both radio frequency (RF) front-ends that are employed in the measurement system are equally constructed as described in this section. Due to the superior performance of substrate integrated technology compared to other planar technology at frequencies higher than 30 GHz [16], substrate integrated waveguide (SIW) has been chosen as main transmission line type. However, interconnections to other transmission line styles are inevitable, as especially active components, e.g., amplifiers and diodes, need to be integrated to the RF structure. The front-end is illustrated as block diagram in Fig.3 and can be divided in several components, which are the signal source, the radar coupler, the antenna, the variable attenuator, the low noise amplifier (LNA) and the six-port junction including the power detectors. The generated signal $S_{LO}$ is equally split up by the radar coupler and half of the power ($S_{Tx}$) is radiated by the antenna. The other part ($S'_2$) is coupled to the reference path where its power level is adjusted using a variable attenuator. The transmitted wave is reflected at the target’s surface and partly received by the antenna structure. An LNA is utilized in order to increase the receive signal...
power ($S'_1$) and to achieve a low total noise figure [7]. Afterward the amplified receive signal $S_1$ and the reference signal $S_2$ are superimposed within the passive six-port junction and down-converted to baseband by four diode detectors [7], [14].

A. Signal source

1) Signal generation: The radio frequency signal source is mainly based on a commercial V-band Infineon BGT60 transceiver millimeter wave integrated circuit (MMIC) providing two balanced signal output pairs for both, the millimeter wave signal (57 GHz to 64 GHz) as well as the reference divider signal. Furthermore, the MMIC features a variable gain amplifier (VGA) enabling an adjustment of the transmit signal power using a serial programming interface (SPI). For settling the frequency of the mm-wave signal, the reference divider of the voltage controlled oscillator (VCO) is connected to a phase-locked-loop (PLL) integrated circuit (IC) of type Hittite HMC703 with an active loop-filter using a differential MSL. The reference clock of 50 MHz for the PLL IC’s phase discriminator is provided by a common temperature controlled Quartz oscillator (TCXO) distributed on the system’s back-end.

2) MMIC-SIW-transition: The connection of the signal source MMIC delivered in an embedded wafer level chip scale package (eWLB) to the passive SIW circuit is accomplished in two steps. First, a two-section half-wave balun in grounded coplanar waveguide (gCPW) technology is used to transform the balanced MMIC output to an unbalanced signal [17], second, a tapered line transition between gCPW and SIW is employed making use of their similar modal fields [18].

B. Radar coupler

Reflecting the work of Henry J. Riblet of 1952 [19], the radar coupler is realized as single layer short slot directional coupler, consisting of two adjacent waveguides with an opening in the common wall [18]. The length of the central aperture which defines the coupling ratio of the directional coupler [20], is chosen for a coupling ratio of -3 dB. In the area of the aperture, the outer waveguide walls are slightly shifted to the middle in order to compensate dispersion effects and enlarging its bandwidth. The exact dimensions of the geometry have been optimized using CST Microwave Studio software.

C. Antenna

As the antenna is forming the interface between radar system and measuring path, an appropriate antenna design suitable for the desired application is essential. By means of the described near-field application, the antenna needs to fulfill several criteria: small aperture size, medium gain with it’s maximum in boresight direction, and easy integration into the chosen highly integrated System-on-Substrate concept with SIW line type on a 127 µm Rogers RO3003 material. In literature, two main SIW antenna types can be identified, which are slotted SIW antennas (e.g., [21], [22]) and leaky-wave SIW antennas (e.g., [23], [24]). However, both types are space-consuming, narrow-band, and their antenna efficiency decreases with reducing substrate height. Furthermore, those antenna types do not meet the requirements of small aperture size. Therefore, other concepts like tapered slot antennas are more suitable in the purposed application [25], [26]. Hence, a SIW surface antenna has been designed based on an end-fire Vivaldi concept which has been first-time presented by the authors in [27],[28] In order to achieve better system integrability and more symmetric elevation behavior, the antenna concept
has been extended by two aluminum shielding caps mounted on both sides of the PCB as shown in Fig. 4. This figure also provides the alignment of the reference coordinate system, which was defined for the calculation of the antennas' farfield data. Fig. 5 depicts a two-dimensional surf plot of the realized antenna gain, providing the gain in dBi as color information. In Fig. 6, the antenna gain is additionally plotted in the xz-plane (delineated in Fig. 4), as this pattern is mainly relevant for the previously described application. For moving targets parallel to the yz-plane (refer to Fig. 4), the antenna pattern introduces a low-pass behavior to the measured distance information. As to be obtained by Fig. 6, the introduced shielding of the antenna structure results in both an increased gain for boresight (ϕ = 90 degrees, ϑ = 0 degrees) and less steep filter slopes of the low-pass behavior.

**D. Low noise amplifier**

The amplification of the received signal is performed by a commercial LNA MMIC of type UMS CHA2159-099F supplied by a dedicated driver network. Since the device is delivered as bare die it needs to be bonded to the substrate material using a cavity in order to keep the wires as short as possible. Fig. 7 gives the layout of the LNA assembly and the designed matching networks for the input and output signal paths referred as S′1 and S1, respectively. An equivalent circuit of the input and output matching and the transition to SIW is displayed in Fig. 8. First, the transformation of the TE10 mode of the SIW to qTEM mode of the conductor backed coplanar waveguide is conducted by a linear tapered microstrip section. The top layer ground conductor is tapered in order to reduce discontinuity effects, additionally. The length of the tapered section is approximately up to quarter wavelength in SIW, the dimensions have been optimized using a full-wave EM simulation tool. The matching of the chip’s input and output impedance \( Z_{in}/Z_{out} \) with the serial ribbon bond wire of inductance \( L_b \) is done by a high impedance transmission line and a parallel capacitance, which is realized as two parallel low impedance open stubs. Therefore, the length of the bond wire is crucial to the input and output matching of the LNA. In order to achieve a good matching and low loss the length of the bond wires has been optimized to be as short as possible (330 µm) by mechanical testing. Furthermore, the length and the shape were considered within the EM simulation of the LNA’s matching network.

**E. Variable attenuator**

In many millimeter wave systems variable attenuators are based on MMIC technology [29], [30], [31]. In general, both transmission type and reflection type attenuators are also realizable using discrete components [32], however, reflection type approaches seem to be more feasible at V-band as the requirement specification concerning the parasitics of the active element is lower [33]. Reconfigurable SIW structures have been shown in several publications many of them dealing with tunable filters [34] or phase shifters [35]. In [36] a reflective varactor tuned SIW phase shifter based on a quadrature hybrid with reflective loads is presented, making use of transforming the TE10 mode of SIW to a qTEM mode of a parallel plate
line. Such design is also feasible for an attenuator structure replacing the varactors by PIN diodes. However, a transition of SIW mode to coplanar waveguide or microstrip line is more straightforward, less space consuming and the design of the biasing network is easier to accomplish.

The proposed reflective attenuator is based on the presented radar coupler structure and two identical reflective loads connected to the coupler’s output ports. The operation principle of suchlike attenuator is extensively described in [33]. The reflective load is displayed in Fig. 9 and is mainly based on a short SIW-to-MSL transformation network and an Aluminum-Gallium-Arsenide (AlGaAs) flip-chip PIN diode of type *Macom MA4AGFCP910*, which is qualified for operation up to more than 70 GHz. In order to reduce the transmission line’s influence, the diode is directly soldered on the tapered section of the transformation network and connected in shunt configuration. The RF ground connection and the bias network are realized using two quarter-wave radial stubs and a high impedance transmission line section of a quarter wavelength. The attenuation value is set by $V_{\text{tune}}$, provided by a digital-to-analog converter (DAC), connected to the diode using a serial resistor limiting the bias current. The attenuation of the described attenuator is adjustable using a 10-bit DAC in a range from 10 dB to 35 dB. While the minimum attenuation value is limited by the series impedance of the PIN diodes, the maximum attenuation value depends on the isolation value of the utilized short-slot coupler.

**F. Six-port receiver**

1) **Passive six-port junction**: Analogous to the six-port designs in microstrip technology, six-port junctions are mostly based on combinations of power dividers, couplers and phase shifters [37]. For the design of single layer six-port junctions in SIW technology two types of power dividers are feasible. Those are the T-junction and the Y-junction [38], [39]. Furthermore, two coupler designs are interesting for a single layer design: The well-known short-slot coupler as well as a cruciform shaped directional coupler [40]. Although the use of cruciform shaped couplers allows for highly space efficient six-port designs [41], its bandwidth as well as its requirement for low manufacturing tolerances plead for utilization of parallel line short-slot couplers. Fig. 10 shows the designed six-port wave correlator structure using three short-slot couplers and one T-junction power divider. In order to achieve high relative bandwidth and stability against manufacturing tolerances, a symmetrical design without the use of linear delay elements has been realized [42], [37]. The input ports have the designators $P_1$ and $P_2$, while $P_3$-$P_6$ are the output ports of the structure. Furthermore, port $P_2$ is terminated by a match, realized using a SIW-MSL transition and a 50Ω flip-chip high frequency thin film resistor of type *Vishay Sfernice CH0201650RGFT* in shunt configuration. By means of the symmetric structure, the amplitude and phase imbalance of the six-port junction is directly related to the imbalance of the designed short-slot couplers. Therefore, the amplitude imbalance $\Delta a_c$, of the designed passive six-port junction is less or equal to 0.6 dB, while the phase imbalance $\Delta \phi_c$ is smaller than 3 degrees over the interesting frequency range.

![Six-port wave correlator structure.](image)

2) **Power detector**: The conversion of the four six-port correlator output signals $S_1$-$S_6$ is achieved by four matched zero bias Schottky diode detectors [14]. The detector structure shown in Fig. 11 is mainly based on a prior design utilizing a low-cost *Skyworks SMS7630-061* silicon zero bias diode, accurately described in [42]. As the influence of parasitic elements gets crucial with increasing frequency, a Gallium-Arsenide (GaAs) zero bias Schottky diode of type *Keysight HSCH9161* in discrete beam lead housing has been chosen for a redesign featuring higher voltage sensitivity and lower noise floor due to its alternative semiconductor technology with lower parasitics. The reactive matching network as well as the output filter have been adapted as shown in Fig. 11.

![Layout of the detector circuit (R=51kΩ, C=12pF).](image)

It is common knowledge and specially shown for six-port receivers in [43], that the ambient temperature has a strong influence on a Schottky diode detector’s characteristic. Thus, temperature effects can’t be neglected in six-port radar systems. Related to the temperature distribution over the complete passive circuit board, a temperature compensation needs to be applied to every single digitized baseband signal within the signal processing routine.
IV. NON-IDEAL BEHAVIOR OF SIX-PORT RADAR SYSTEMS

Non-ideal effects degrade the Six-Port radar system’s performance [7]. In general, there are three non-ideal effects to be identified, which are noise, limited isolation and the receiver non-linearity [44], which is mainly based on the non-ideal detector behavior. The entire described non-ideal behavior can be categorized in systematic and random effects. As random noise effects introduced by the signal source and the receiver structure have little impact on short-range systems and are easy to overcome, the primary concern in the presented CW (Six-Port) radar system is the minimization of the systematic measurement error [45], which results from limited isolation (crosstalk) between the transmit (Tx) and the receive (Rx) path and detector non-linearity.

A. Influence of impairments

Fig. 3 showed the complete frontend of the radar system with the passive six-port junction. In this section non-idealities of the proposed RF frontend will be identified and their influence on the baseband signals will be evaluated by an analytic derivation of the six-port equations with a crosstalk signal. The reasons for limited isolation between the transmit and the receive paths are manifold. Parasitic coupling of the oscillator signal to the receiver structure occurs due to limited shielding. Limited isolation of the radar coupler (\(S_{\text{leakage}}\)) antenna mismatch (\(S_{\text{antenna}}\)), and reflections of static targets in the antenna’s field-of-view (\(S_{\text{Rx,par}}\)) [46], [47]. Therefore, the sum signal \(S_{\text{Rx,par}}\) at the receiver’s input port is composed as follows:

\[
S_{\text{Rx,par}} = S_{\text{Rx}} + S_{\text{leakage}} + S_{\text{antenna}} + \sum S_{\text{Rx,par}}. \tag{1}
\]

Thus, the crosstalk signal \(S_X\) can be expressed by the sum of all parasitic signal parts:

\[
S_X = S_{\text{leakage}} + S_{\text{antenna}} + \sum S_{\text{Rx,par}}. \tag{2}
\]

As the parasitic signals directly interfere with the receive signal, the LNA and the reconfigurable attenuator must adjust the power level of the reference signal \(S_2 = A_2 e^{j\omega_2 t + \phi_2}\) and the receive signal \(S_1 = A_1 e^{j\omega_1 t + \phi_1}\) to optimize the baseband voltages so that the dynamic range of the baseband circuitry can be efficiently used.

To analyze the influence of parasitic signals in the receive path and to find the optimal power level settings, the six-port equations [48] with an additional parasitic signal \(S_X = A_X e^{j\omega_X t + \phi_X}\) need to be calculated for the homodyne architecture (\(\omega_1 = \omega_2 = \omega_X\)):

\[
B_3/k_3 = A_1^2 + A_2^2 + 2A_1A_2 \sin(\phi_1 - \phi_2) + A_X^2 + 2A_2 A_X \sin(\phi_X - \phi_2) + 2A_1 A_X \cos(\phi_1 - \phi_X), \tag{3}
\]

\[
B_4/k_4 = A_1^2 + A_2^2 + 2A_1A_2 \sin(\phi_1 - \phi_2) + A_X^2 - 2A_2 A_X \sin(\phi_X - \phi_2) + 2A_1 A_X \cos(\phi_1 - \phi_X), \tag{4}
\]

\[
B_5/k_5 = A_1^2 + A_2^2 + 2A_1A_2 \cos(\phi_1 - \phi_2) + A_X^2 + 2A_2 A_X \cos(\phi_X - \phi_2) + 2A_1 A_X \cos(\phi_1 - \phi_X), \tag{5}
\]

\[
B_6/k_6 = A_1^2 + A_2^2 - 2A_1A_2 \cos(\phi_1 - \phi_2) + A_X^2 - 2A_2 A_X \cos(\phi_X - \phi_2) + 2A_1 A_X \cos(\phi_1 - \phi_X), \tag{6}
\]

where \(k_i\) are conversion factors of the four power detectors.

When analyzing the obtained baseband signals, each signal again can be split into two parts: the wanted signal that is independent from the crosstalk signal, and the unwanted signal that depends on the crosstalk signal’s amplitude and phase. The unwanted signal itself consists of three components:

1) an offset equal to the square of the crosstalk signal’s amplitude
2) an offset proportional to the amplitudes of the reference and the crosstalk signal and dependent on the phase difference between these two signals (note, that \(\phi_2\) and \(\phi_X\) are constant)
3) a sinusoidal signal that varies with the rotating phase \(\phi_1\) (note, that the phase of the receive signal is varying with the target’s movement)

For a moving target, the wanted signal and the third component of the unwanted signal vary with the same rotating phase \(\phi_1(t)\). Thus, the sum of both signals also varies with the rotating phase \(\phi_1(t)\), but has a different phase offset and amplitude. For instance, the sum of the varying part of baseband voltage \(B_3\) can be calculated to be:

\[
2A_1A_2 \sin(\phi_1 - \phi_2) + 2A_1 A_X \cos(\phi_1 - \phi_X) = A_x \sin(\phi_2 + \phi_s) \tag{7}
\]

with

\[
A_x = 2A_1 \sqrt{A_2^2 + A_X^2 + 2A_2 A_X \sin(\phi_X - \phi_2)}, \tag{8}
\]

\[
\phi_s = \arccos \left( \frac{A_2 \cos(\phi_2) + A_X \sin(\phi_X)}{\sqrt{A_2^2 + A_X^2 + 2A_2 A_X \sin(\phi_X - \phi_2)}} \right). \tag{9}
\]

This demonstrates, that a crosstalk signal results in three different types of errors:

1) Offset: even if ideal detectors are supposed with \(k_3 = k_4 = k_5 = k_6\), the offsets in (3) to (6) do not fully cancel out by calculating the IQ signals, but an offset \(4k_2 A_2 A_X \sin(\phi_X - \phi_2)\) and \(4k_2 A_2 A_X \cos(\phi_X - \phi_2)\), respectively, is remaining.

2) Amplitude imbalance: the amplitudes of the baseband signals do not only depend on the amplitude of the reference and the received signal, but also on the crosstalk signal’s amplitude and the phase shift between reference and crosstalk signal.

3) Phase imbalance: as to be obtained by (7) and (9) there is also a phase imbalance that has to be considered. This means that the four baseband signals are not separated by 90° anymore.

These errors usually can be eliminated with an IQ imbalance compensation [49]. However, due to the characteristics of the antenna, the amplitude \(A_1\) of the received signal is exponentially decaying with increasing distance to the target. Thus, the baseband signals show, additionally to the above mentioned errors, a spiral characteristic with a decaying diameter over the distance in the near field region [50].
B. Influence of the Six-Port receiver’s non-linearity

The basic concept of Six-Port technology in measurement applications is mainly based on two requirements, which are defined relative phase shifts between the detector input signals and a known transfer function of the used power detectors, providing square-law behavior. Deviations will affect the obtained baseband signals and, therefore, the value of the complex signal. Consequently, receiver non-linearity can be directly assigned to the non-linearity of the detector characteristics. However, the amplitude and phase imbalances $\Delta \alpha_c$ and $\Delta \phi_c$ of the passive Six-Port junction, which can be totally compensated for an ideal detector characteristic, increase the non-ideal behavior of the receiver. While amplitude and phase imbalances $\Delta \alpha_c$ and $\Delta \phi_c$ can be addressed by a proper design of the used couplers of the Six-Port junction, the influence of the four detector transfer functions is more complex by means of different reasons: Each detector diode is subject to individual tolerances in placement as well as in the semiconductor manufacturing process. Therefore, each of the four detector devices has an individual transfer function. Due to the necessary high dynamic range as well as the related power imbalance deviation from square law occurs by reasons of the diodes’ non-linearity [51]. The effect of non-linear detector characteristics on the IQ data was analyzed for constant power ratios in [44] and a strong deformation of the ellipse could be determined at high input powers. However, in the present application each detector runs at different input power levels.

\[ \begin{align*}
(a) & \quad \text{IQ (square-law detectors)} \\
(b) & \quad \text{IQ (non-linear detectors)}
\end{align*} \]

Fig. 12. Calculated baseband signals in the IQ representation using (a) square-law detector characteristic and (b) non-linear characteristic.

The impact of the described hardware deviations on the offset compensated simulated IQ-data of the application is displayed in Fig. 12. The presented data are based on a detailed system simulation, considering the described effects of impairments, power imbalance, amplitude and phase imbalances, and detector non-linearity. The simulation data are generated for an increasing distance sweep of 5 mm range, starting at 22.5 mm. The antenna is simulated with an ideal pattern providing 8 dBi. Multiple reflections between antenna and target are not considered in the system simulation. Fig. 12(a) shows the obtained offset-compensated spiral for square law detectors, while Fig. 12(b) is based on a non-linear detector behavior. It can be obtained, that the second spiral shows deformations, which can be related to the changing input power level.

V. NF Backend

The backend PCB contains all components for voltage and control signal generation, the analog baseband processing, as well as signal processing and communication. The board consists of six layers and its U-shape is optimized to align the antennas of the RF front-end modules directly opposite in a distance of 5 cm (1.97 inch) including a cut-out for the metal sheet, as depicted in Fig. 13b.

A. Analog signal conditioning and digitization

The analog baseband consists of an analog signal conditioning stage with digital adjustable gain and an 8-channel simultaneous sampling ADC. At first a low-pass filtering and amplification of the susceptible detector output signals is performed by THS4524 operational amplifiers from Texas Instruments (TI). Furthermore, a digitally adjustable potentiometer MCP41-series from Microchip is used to realize variable gain amplifiers with a run-time adjustable gain up to 255. Based on the design considerations from [52] an 8-channel simultaneous sampling 24-bit ADC (ADS1278 from TI) has been selected. Due to the Delta-Sigma ($\Delta \Sigma$) architecture the filter requirements of the previous anti-aliasing stage can be relaxed while still maintaining a fast sampling rate of up to 105 kSa/s that is more than sufficient even for fast moving targets.

B. Microcontroller and communication

For controlling and monitoring of the system an Infineon XMC4500 microcontroller (MCU) is used, which can be accessed by an 100-Mbps-ethernet-link to change the system parameters, i.e., the VCO output frequencies, the output power of the RF modules, and the baseband amplification, using several SPI connections. Furthermore, the MCU reads the data from the ADC’s output registers and sends the raw data as well as some monitoring signals, e.g., the temperatures of the RF MMICs, to a personal computer, which performs the further signal processing.

VI. Calibration and Measurement

Testing the sensor in a real roll milling application is both complicated and not mandatory for initial testing and review of the proposed sensor. Therefore, the evaluation of the sensor’s measurement principle is based on a simplified measurement setup, which also can be undertaken under laboratory conditions.

A. Measurement Setup

The measurement setup mainly consists of the previously described hardware and an aluminum plate which is mounted on two orthogonal linear stages. A photograph of the complete setup is depicted in Fig. 13a. The schematic top view of this setup is shown in Fig. 13b, illustrating the range directions (X-Range and Y-Range) and the distances between antennas and target ($d_A$ and $d_B$). Here, the two front-ends are named FE-A and FE-B, respectively. The orthogonal positioned linear stages can be moved with an accuracy of about 500 nm.
The signal curve is an elliptic appearance of the before mentioned constant offset and phase imbalances. The resulting complex data with the following formula:

\[ IQ_\eta = (B_{\eta,5} - B_{\eta,6}) + j(B_{\eta,3} - B_{\eta,4}) \]  

Since these values are impaired by offsets and phase imbalances, the center of the measured spiral is estimated by using the technique introduced in [50]. The offset-compensated ellipse, denoted as \( IQ_{\text{cal},\eta,i} \), is shown in Fig. 14(c) and (d). The remaining phase non-linearity is illustrated in Fig. 14(e) and (f). To compensate these phase imbalances, the measured phase values are mapped on the ideal phase values, which are derived from the positions of Stage-X by means of

\[ \phi_{\text{ideal},\eta} = \frac{4\pi}{\lambda} x, \]

wherein \( d_{\text{ideal},\eta} \) denotes the ideal distances between front-end antennas and the target and \( \lambda \) denotes the wavelength of the emitted electromagnetic wave. The ideal phase combined with the measured phase can subsequently be used to correct the phase during actual measurements which are described in the following subsection.

B. Calibration

Moving the plate in X-Range ideally results in a linear phase shift. Due to the non-idealities in the RF front-ends presented in Sec. IV, the measured quantity values are impaired by a constant offset and phase imbalances. The resulting complex baseband signals can be calculated from

\[ \eta = \begin{bmatrix} \eta_1 \\ \eta_2 \\ \eta_3 \\ \eta_4 \end{bmatrix}, \]

These voltages can be used to calculate the complex data \( IQ_\eta \) for each front-end \( \eta \) data with the following formula:

The aluminum plate is shaped with five steps, each having a height of approximately 100 µm (≈ 3.937 mil) and a width of about 5 cm (≈ 1.96 inch). The structured side of the plate is facing FE-A while the opposite surface is completely flat. The manufacturing quality of the plate corresponds to the precision of the employed manufacturing computer numerically controlled milling process, which is 10 µm (≈ 0.4 mil) in all directions.

(≈ 0.2 mil). Both are controlled by a computer which also communicates with the microcontroller. The stage moving in X-Range is named Stage-X in the following, while the other one is termed as Stage-Y.

The two radar systems are placed on the back side of the U-shaped back-end PCB.
C. Measurements

The complete signal processing flow of the actual measurements is shown in Fig. 15, in which the first four steps are equal to the calibration steps. The offset compensation of the measurement signal is accomplished by subtracting the offset value determined by the prior calibration routine. The resulting IQ signals are denoted as $IQ_{\text{comp},n}[\eta]$. The resulting phase non-linearity due to the elliptic distortion is compensated by a phase error correction. In general, there are several ways to realize such a phase error correction. One possibility is a big look-up-table with a lot of calibration points which implies a huge calibration effort and requires a lot of memory space. Another approach is to significantly reduce the number of calibration points and use a segmental polynomial approximation for interpolation [55]. In this measurement scenario a combination of a look-up-table and a spline-interpolation is utilized.

Fig. 14 (e) and (f) show that the measured phase values $\phi_{\text{meas},n}$ can be mapped on the ideal phase values $\phi_{\text{ideal},n}$ along the X-Range by a bijective function $\Phi$. This function can be obtained at least point-wise from the calibration data while continuity is realized by spline interpolation. The inversion of the function, denoted as $\Phi^{-1}$, can be employed to obtain the corrected phase values $\phi_{\text{corr},n}$:

$$\phi_{\text{cal},n} = \Phi(\phi_{\text{ideal},n})$$
$$\phi_{\text{corr},n} = \Phi^{-1}(\arg \{ IQ_{\text{comp},n} \})$$

Comparing the baseband signals of both frontends, which are depicted in Fig. 14(a),(b), the detectors are used in diverse operating points which leads to differently distorted IQ signals. Since the magnitude distortion can be expressed as a function of the phase it is possible to correct the magnitude errors with:

$$| IQ_{\text{corr},n} | = \frac{| IQ_{\text{comp},n} |}{| IQ_{\text{cal},n} | (\arg \{ IQ_{\text{comp},n} \})}$$
$$IQ_{\text{corr},n} = | IQ_{\text{corr},n} | \cdot e^{j \phi_{\text{corr},n}}.$$  

Fig. 16 illustrates the magnitudes of the calibration data as well as the offset-compensated output signal both of FE-B. The amplitude distortions of the measurement data in subplot (b) can be reduced by incorporating the knowledge from the calibration measurements depicted in subplot (a) using (14). Only phase information is needed to calculate the relative distances. The magnitude is linearized to get the valid complex signal $IQ_{\text{corr},n}$. A low-pass filter is applied to reduce the influence of noise. To get a reasonable cut-off frequency for this filter, the spectrum of $IQ_{\text{corr},n}$ has to be considered, which is determined by two factors. The first factor is the
surface of the plate, which is unknown a-priori, while the second factor is the antenna pattern. A wider antenna pattern applies a stronger low-pass effect on the measured data. This influence can be described quantitatively by convolving the complex surface reflection function with a convolution kernel, that incorporates the antenna pattern as well as reflections in the near-field. The convolution kernel is determined by simulation and is denoted as \( \theta \). The low-pass effect can now be described quantitatively by considering the Fourier transform of the convolution kernel. A Bessel filter is applied on the corrected IQ-data due to its constant group delay. The cutoff frequency \( \omega_{cutoff} \) is chosen as

\[
40 \text{ dB} = |\mathcal{F}\{\theta\}(0)|_{\text{dB}} - |\mathcal{F}\{\theta\}(\omega_{cutoff})|_{\text{dB}}.
\]  

A Bessel filter with the cutoff frequency \( \omega_{cutoff} \) and a filter order of 15 delivers an appropriate performance in terms of noise reduction combined with insignificant distortions of the phase data.

The filtered IQ data signal is used to calculate the phase data by means of \( \phi = \angle IQ \). Regarding (11) the relative distances which are illustrated in Fig. 17 for the target moving 15 cm (5.9 inch) along the Y-Range can finally be computed.

The steps on the side of the aluminum plate facing FE-A can be recognized. Since the plate is not aligned perfectly orthogonal to the antenna beams, linear distance changes are detected along the Y-Range by both frontends. The relative thickness of the plate is calculated utilizing the calculated relative distances of both front-ends:

\[
D = -(d_A + d_B).
\]  

The relative thickness of the aluminum plate along a Y-Range of 15 cm is depicted in Fig. 18. As a result of the differential measurement principle, the linear distance changes are canceled out, here. Furthermore, the figure depicts the expected target surface with steps of a height of 110 \( \mu \)m and a size of 5 cm on one side of the aluminum plate. The decreased steepness of the steps is caused by the low-pass effect mentioned above, while the ripple on the obtained thickness curve can be directly related to the convolution of the antenna diagram and the steps of the surface. In a sheet thickness application such ripple will not occur, as thickness changes are expected to be smooth and frequency dependent reflection effects with the exception of multiple reflections between metal sheet and antenna are not present. Slow thickness drifts, however, will always be obtained by the measurement principle. For distance variations limited to a quarter wavelength between two measurement values, the implemented algorithm is able to follow, even if both values are in different unambiguous ranges. However, such immediate changes will only occur during serious error events.

VII. COMPARISON WITH STATE-OF-THE-ART

There is a wide range of different measurement principles being used in industrial applications ranging from ultrasonic to radiography. However, only a few of them meet the demanding requirements of accurately measuring sheet thickness in the described environment. Table I gives an overview of contactless state-of-the-art metrology systems, which are in principle capable of the sheet thickness measurement application, or systems which are specially designed with focus on this intended purpose. Therefore, the schedule is divided in state-of-the-art commercial products for sheet thickness measurement and radar systems which have not been considered for being used in this application, yet.

The existing difficulties of the application have been extensively described in Section I and II. Considering those, the systems need to be evaluated by their measurement range, their precision, their update rate, and the spot diameter of the area under observation. As being related to the wavelength, optical systems show best behavior in terms of precision, update rate as well as the size of the monitored spot. Optical systems also allow for an extraordinary accurate scanning of the surface profiles of fast moving sheets. However, their usability is limited due to well known facts. First of all, they suffer from environmental conditions, such as dust and fog as well as changing illumination. Furthermore, they hardly can monitor glowing surfaces. For those reasons, optical systems are mainly applied in cold mills, where mechanical gauges are also a good alternative. X-ray based systems can more or less be used independent of the surrounding environment. However, for good results the strip’s material compound as well as its temperature need to be known precisely, as the obtained thickness value is directly related to the material absorption coefficient. By nature, noise is also an important issue for X-ray gauges [3] as it is influenced by several parameters which are the air gap between transmitter and detector, the source power and plate current of the tube, the
detector area and finally the material-under-test itself [5]. In order to avoid high statistical noise, the parameters of X-ray gauges need to be adapted for each material change which is a major drawback to the system. Finally, the lifetime of the tubes is strictly limited. Therefore, X-ray detectors require extensive maintenance.

Radar based systems do not suffer from previously described drawbacks and therefore, represent an appropriate alternative. In comparison to optical systems, the wavelength is larger thereby, also the diameter of the monitored spot. An approximation of the spot diameter $d_{3\text{-}dB}$ is given by:

$$d_{3\text{-}dB} = d \arctan(\alpha_{3\text{-}dB})$$  \hspace{1cm} (18)

with $\alpha_{3\text{-}dB}$ being the antenna’s 3-dB beamwidth and $d$ being the propagation distance. However, some part of the power is radiated and received outside the 3-dB spot and the applicability of (18) raises with higher antenna directivity.

FMCW systems are not suitable for the application due to the high required bandwidth, which increases the ramp duration and therefore, decreases the update rate. However, a high update rate is mandatory for the application. Furthermore, bandwidth is strictly limited by regulatory authorities (e.g., ITU), especially as no encapsulated operation is provided. However, single frequency CW radar meets the requirements with an unambiguous range limited to half its wavelength and its update rate only depends on the speed of the ADC, the signal processing, and the speed of communication interface. While the 24 GHz six-port radar system presented in [56] does not show an acceptable update rate as well as lower precision, the proposed 61 GHz radar system is capable of update rates of more than 100 kSa/s, while providing 10 µm precision.

By means of the differential measurement principle, the system is capable of compensating misalignment, while keeping scalability to varying sheet thickness. The measurements presented in Section VI show a low-pass effect resulting from the steps in the measured target. The low-pass and convolution effect will not degrade suitability to the application as on the one hand smooth changes should be monitored and on the other hand rapid steps need to be detected immediately as they are an indicator for a massive failure of the rolling mill.

VIII. CONCLUSION

A new concept for high-speed measurement of sheet-thickness in rolling mills based on two six-port interferometric radar systems in a differential configuration has been presented. The proposed system design is based on a detailed analysis of the special demands resulting from the application. The radio frequency frontends are realized as system on substrate in a mixed SIW-MSL-technology to make the systems insensitive to external interference. SIW technology enables an encapsulated frontend design allowing for an appropriate antenna design as well as enabling a good thermal design.

Furthermore, a detailed analysis of occuring non-ideal effects of the radio-frequency frontend was provided and a dedicated calibration strategy has been presented. The validation of the proposed measurement system has been performed under laboratory conditions delivering promising results for the desired application.

A discussion and comparison with present state-of-the-art indicates, that the millimeter-wave interferometric measurement principle is a promising alternative to current technology. Especially the suitability to the demanding environmental conditions and high update rates as well as the need for comparably low maintenance are factors that underline the significant difference to currently used technology.

REFERENCES


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