In-field Receiver and Processing Systems for MRI

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Summary

For better image quality, faster imaging procedures and more comfortable acquisition, magnetic resonance imaging (MRI) is evolving into a multi-channel and multi-modal instrument. Beyond the current state-of-the-art, where high channel count coil arrays already notably reduce scan times and increase sensitivity, future high precision magnetic resonance (MR) imaging will not only make wide use of flexible and wearable coil arrays, but rely specifically on information from a multitude of ancillary sensors. Relevant sensor data will be smartly combined in real-time to correct scanner imperfections and patient motion, to analyze and act upon physiological information, and to enable real-time interventional imaging.

Today, imaging coil arrays and the wide adoption of auxiliary sensors are hitting a limit in terms of handling and mechanical dimensions, as well as signal integrity and safety. This is largely due to the required cable installations that are costly and ponderous. An alternative to the large number of necessary radio frequency (RF) cables is needed. For this purpose, in this work, it is proposed to operate the sensitive MRI receive electronics in-bore, including digitization and signal pre-processing. Cables can be avoided if the spectrometer is placed as close to the signal source as possible and signal transmission is realized on an alternative medium, e.g. wireless or optical links.

Therefore, an in-bore receiver for imaging and field monitoring, capable of operation in the adverse environment of the MRI scanner bore, was developed and evaluated. Potential vulnerabilities were identified and
limitations and design principles for electronics for in-bore use were explored. Signal integrity, scalability and handling during operation, without compromising bandwidth or acquisition duty cycle, were key considerations for building a 16-channel receiver module. The compact receiver allows for broadband, high dynamic range signal acquisition and was designed to be modular for easy adaptation, including the acquisition of different nuclei. A powerful field-programmable gate array (FPGA) enables pre-processing of acquired data before it is transmitted by a single, non-galvanic high-speed optical link.

It is shown that multi-channel spectrometers can be built and operated in-bore and that high quality signal digitization is possible even in the harsh environment of the MR scanner bore. Long and coupling cables were replaced by a single optical link and the performance of the constructed system is on-par with commercial out-bore receivers. It even allows for sensitive field monitoring. In special cases, scanner electromagnetic interference (EMI), especially gradient action, can hamper in-bore clocking and hence acquisition performance. To mitigate this, a solution was presented to improve overall signal-to-noise ratio (SNR) based on clock stabilization using feedback.

The receiver is unique in MRI in that it allows seamless acquisition of multi-nuclear signals. Its modularity and the re-configurable digital subsystem make it a convenient platform for the development and integration of other sensor and actuator frontends. The flexibility of the frontends and the versatile digital processing pipeline has many applications of which a few were demonstrated.

When scaling the number of receive channels, the vast amount of data collected is becoming a limitation. Data from imaging coils and sensors somehow needs to be processed, stored, and conceivably acted upon. Thus, data compression, early pre-processing, and data interpretation with low latency are becoming ever more important. For example, the required bandwidth of the effectively used imaging trajectory is significantly smaller than the recorded data from typically 16 or more field probes. Hence, if one calculates the trajectory on the hardware, massive amounts of memory and calculation time can be saved.
To address these digitization and processing challenges and to enable both enhanced image reconstruction based on sensor information and run-time updates of scanner operation, an acquisition and processing platform was developed and presented. The built out-field processing system features generic high throughput digital interfaces and was demonstrated to connect multiple in-bore operating receivers which deliver heterogeneous types of data. Its highly stable reference clock ensures synchronous acquisition of data from all connected modules. This is crucial for e.g. retrospective correction of imaging data based on field probe trajectory measurements. It is designed to be daisy-chained for ultimate scalability and large numbers of channels. At the heart of the system sits a System-on-Chip (SoC), a combination of dual-core processor and programmable logic, which offers ample resources for multi-channel parallel signal processing. The realized processing pipeline and application architecture allows on-line adjustment of parameters of the various implemented computation blocks (including matrix-vector multiplier, COordinate Rotation Digital Computer (CORDIC), filters, etc.) and the signal routing among them. The design is flexible enough to allow integration of further processing blocks with minimal effort.

Several of the processing steps needed in MRI can be moved to hardware. As an example, coil compression is a first reconstruction step that lowers the computational burden and data at the cost of minor SNR loss. Similar is trajectory extraction from field probes. Enabled by the novel platform, fully synchronous acquisition of imaging and field monitoring data from multiple receivers was demonstrated, allowing seamless reconstruction of even the most demanding sequences. The power of simultaneous real-time k-space trajectory extraction and coil compression in hardware was illustrated, which resulted in significant data savings without loss of flexibility and image quality.

Besides the discussed cabling and processing challenges, MRI coils suffer from a lack of patient comfort and parallel imaging methods can not be used to their full potential due to the absence of arrays that tightly conform to the patient anatomy. Likewise, mechanical rigidity prevents changes in posture, such as flexion of joints, needed for new functional anatomical insights. This defiance can be addressed by building flexible, wearable and digital detector arrays.
The know-how gained in operating receive electronics in-bore and the joined forces of three institutes of ETH Zürich has allowed the integration and operation of a custom designed MRI receiver chip, capable of digitizing the imaging data directly on-coil. In principle the whole receiver chain can be integrated on a single chip, enabling wearable multi-channel coil arrays.

As proof of concept, a wearable, flexible four channel receive array was presented and successful imaging on the human wrist was demonstrated. The custom integration of an MRI receiver chip is a radical change of how MR signals are acquired. The introduction of an on-coil MR receiver enables better exploitation of the power of receive array coils and is an essential step towards replacing the cage-like detectors used in today’s clinical MRI for improved patient comfort, sensitivity, and new insights enabled by the ability to change posture.

To conclude, MRI acquisition and processing hardware that surpasses the abilities of existing platforms and introduces the paradigm of distributed computing to MR, is presented. We went from out-field to in-bore data acquisition, from offline to large scale real-time processing, and finally even to on-body digitization.
Zusammenfassung


Sowohl Spulen als auch Sensoren, die in der Magnetröhre ihre Anwendung finden, werden traditionell mit langen Kabeln zum Empfänger ausserhalb des MRT verbunden. Diese teuren und schwerfälligen Kabelinstallationen sind ein Problem für die Entwicklung von noch spezifischeren Spulenarrays und halten die weite Verbreitung von Hilfssensoren auf. Nicht nur die Handhabung und die physikalische Abmessung der Kabel spielt dabei eine wichtige Rolle, sondern auch die mit den zahlreichen


In der Arbeit wird ebenfalls darauf eingegangen, dass in besonderen Fällen die elektromagnetische Umgebung im Scanner die Akquisitionsleistung eines Empfängers beeinflussen kann, insbesondere das verletzliche
Zeitsignal zur Digitalisierung. Auch dafür wird eine Lösung vorgestellt, die mittels Rückkopplung negative Einflüsse kompensiert.


Einige der erforderlichen Verarbeitungsschritte im Bildgebungsprozess von MRT können auf die Hardware übertragen werden. Ein Beispiel dafür ist die Echtzeit-Spulenkompression als ein erster Rekonstruktionsschritt, die den Rechenaufwand und die Daten auf Kosten eines sehr geringen SNR-Verlusts senkt. Ähnlich ist die Gradiententrajektorienberechnung aus Daten von Feldsensoren. In dieser Arbeit wird die absolut synchrone Erfassung von Bild- und Feldsensordaten von mehreren Empfängern
demonstriert, was eine problemlose Rekonstruktion von anspruchsvoller-
ten Sequenzen ermöglicht. Die Leistungsfähigkeit des Systems wurde
an der Echtzeitberechnung der Gradiententrajektorie und simultaner
Spulenkompression in Hardware demonstriert. Ein großer Teil der Daten
kann damit eingespart werden, ohne Einschränkung der Flexibilität oder
eine Reduktion der Bildqualität hinnehmen zu müssen.

Nebst den erwähnten Kabel- und Verarbeitungsherausforderungen,
sind MRT-Spulen traditionell sehr unkomfortabel für Patienten. Die volle
Leistung paralleler Bildgebung zur Scanbeschleunigung kann aufgrund
des Fehlens von Arrays, die eng an die Anatomie des Patienten an-
gepasst sind, nicht voll ausgeschöpft werden. Die starren Spulenarrays
verhindern zudem eine Veränderungen der Körperhaltung, zum Beispiel
das Beugen der Gelenke, die für funktionelle anatomische Einsichten
von grossem Wert sind. Dieser Herausforderung kann begegnet werden,
indem flexible, tragbare und digitale Detektor-Arrays gebaut werden.

Das Know-how zum Betrieb von Empfangselektronik im Magnetfeld und
die Zusammenarbeit von drei Instituten der ETH Zürich ermöglichte die
Integration und den Betrieb eines massgeschneiderten Empfängerschips,
der die Bilddaten direkt auf der Spule digitalisieren kann. Die gesamte
Empfangskette wird auf einem einzigen Chip integriert und Spulenarrays
die wie ein Kleidungsstück tragbar sind, können gebaut werden.

In einem ersten Schritt wurde ein tragbares, flexibles vierkanal Emp-
fangsarray vorgestellt und eine erfolgreiche Bildgebung am menschlichen
Handgelenk demonstriert. Die Integration eines MR-Empfängerschips
ist eine radikale Änderung in der Art und Weise, wie MR-Signale erfasst
werden. Dieser On-Coil-MR-Empfänger ermöglicht eine bessere Aus-
nutzung der Leistung von Empfangsarrayspulen und ist ein wesentlicher
Schritt zum Ersetzen der käfigartigen Detektoren, wie sie in der heutigen
klinischen MRT Verwendung finden.

Kurzum, diese Arbeit stellt neuartige MRT Empfangs- und Verarbei-
tungshardware vor, die die Fähigkeiten bestehender Plattformen übertrifft
und das Paradigma des verteilten Rechnens in MR einführt. Dafür haben
wir den Empfänger in die Magnetöhre gebracht, sind von Offline- zu
Echtzeitverarbeitung übergegangen und schliesslich sogar zur Signalauf-
nahme direkt auf dem Körper eines Patienten.
for Julia and Rahel
- my beacons of light
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Chapter 1

Introduction

Linking some profoundly interesting principles of physics with highly sophisticated engineering, MRI has matured to be a widely used imaging modality in day-to-day medical diagnostics and leading-edge research.

MRI owes its success to the ability to depict a large variety of anatomical and functional features, processes and diseases at salient spatial resolution and with high diagnostic accuracy in 3D. Unmatched by other tomographic methods that often rely on ionizing radiation or radioactive tracers, MRI is considered innocuous. It builds on the fundamental characteristic of nuclei to absorb and re-emit electromagnetic energy when placed in a magnetic field. Clever manipulation of both the magnetic field and the energy delivered to the sample, paired with delicate sensing, reveals otherwise inaccessible physical and chemical properties. With this, and in contrast to other medical imaging methods that rely on energy scattering, MRI achieves high-resolution images even though the wavelength of the electromagnetic energy used is much larger than the features observed. Good reasons for some to call it the 'jewel in the crown' of medical technology [1].

Albeit MRI grants deep insights, present-day scanners tend to face limitations in terms of SNR and are commonly known for lengthy and
costly scan sessions. To remedy these detriments, many technical advancements have already been introduced to MRI. Among others, it was found that by using multiple receive coils scanning can significantly be accelerated, that additional sensors can preempt limited gradient fidelity or help to correct for patient motion, and that by real-time sensing and actuation several improvements to the homogeneity of involved electromagnetic fields can be achieved. However, these methods demand for ever increasing numbers of receive channels of diverse types. More and more cabling is required to connect sensors and receive coils, hitting limits in terms of handling and mechanical dimensions, as well as signal integrity and safety due to RF coupling. Not to mention the vast amount of data that needs to be processed and stored.

Besides, MRI scans suffer from lack of patient comfort. The small scanner bore and the confining, rigid coil arrays mounted around the patient are often enough reason to cause anxiety. Patient setup time takes up a significant amount of total exam time, proliferating cost.

This thesis explores the possibilities of improving some of these impediments by taking on MRI receiver hardware - receiver electronics in particular. Coil arrays can be scaled further if cabling is removed. Data rates can be reduced and sensor data fused by real-time hardware processing. Patient comfort and sensitivity can be improved by building flexible, wearable and digital detector arrays.

For readers not familiar with MRI, the following section presents a short overview on the fundamentals of MRI. Subsequently, a more detailed description of the state-of-the-art, the challenges and the motivation for this work is given. Finally, an overview of the contributions to solving some of the current limitations and an outline of this document conclude the current chapter.
1.1 MRI fundamentals

The noninvasive contrast versatility of MRI and the fact that patients are not exposed to ionizing radiation are made possible by mechanisms depending purely on magnetic fields combined with electromagnetic energy transmission and reception. MRI uses the combination of three intricately choreographed electromagnetic fields for image generation:

A strong static magnetic field ($B_0$) establishes the polarization of atomic nuclei with a dipole moment and leads to their precession about the magnetic background field vector. Several atoms with a dipole moment exist. Among those, hydrogen is the simplest nucleus and the one with the strongest magnetic dipole moment. Since the human body is largely made up from water and fat, both of them containing hydrogen atoms, the most common nuclei for MRI are the hydrogen ($^1H$) protons. Magnetic moments (or spins) of hydrogen protons brought into a magnetic field align either parallel or anti-parallel to this field. Both states occupy different energy levels. Dipoles aligned parallel to the external field reside at a lower energy level than those oriented anti-parallel. Only slightly more spins point parallel to the static field. At human body temperature ($310^\circ$ Kelvin) and a $B_0$ of 1 Tesla solely about three in a million. Boltzmann statistics govern the population difference which is predicted as $\frac{N_+}{N_-} = \exp(-\Delta E/kT)$, where $N_+$ and $N_-$ represent the number of spins aligned parallel or anti-parallel, $\Delta E$ the energy difference, $T$ the absolute Temperature in °K and $k$ the Boltzmann constant.

Being a quantum mechanical phenomenon on a microscopic level, the collective effect of the huge number of nuclei in a sample produces a macroscopic net magnetization ($M$) which can be analyzed using classical physics. For example, a cubic millimeter of tissue contains approximately $10^{19}$ hydrogen atoms. This observable net magnetization, or the energy difference $\Delta E$, depends on the external field $B_0$ as well as the nucleus' magnetic dipole strength $\mu$: $\Delta E = \mu B_0$. However, an established net magnetization does not yet give rise to a signal.

With the help of an oscillating radio frequency field ($B_1$), the polarized spin system can be manipulated. By supplying energy in the
exact amount of the difference of the two energy levels produced by
the external field, dipoles are forced to "flip" from the lower to the
higher energy state. This excitation leads to a tilting of the rotating
macroscopic net magnetization at an angle away from the background
field vector. The net magnetization is resolved into longitudinal ($M_z$)
and transverse components ($M_{xy}$) relative to $B_0$. Before excitation,
the net magnetization is aligned with $B_0$ and hence, $M_z = M_0$ and $M_{xy} = 0$.

The electromagnetic wave providing the energy ($\Delta E$) needed ¹ to excite
the magnetic dipoles is referred to as precession or Larmor frequency
$\omega_0$. From $\Delta E = h\omega_0/(2\pi) = \mu B_0$ and with $\gamma_p = 2\pi \mu_p/h$, the Larmor
equation holds:

$$\omega_0 = \gamma_p B_0.$$  (1.1)

$\gamma_p$ is the gyromagnetic ratio of a particle or nucleus in units of radians
per second per Tesla, i.e. the ratio of the magnetic dipole moment to
its spin angular momentum. For hydrogen the gyromagnetic ratio is
$\gamma_p/(2/\pi) = 42.578 \text{ MHz/T}$.

In MRI the excitation signal is applied using tuned antennas placed in
proximity of the anatomy to be examined. The amount of tipping, the
flip angle $\theta$, is defined by the duration ($T_{ex}$) and strength of the radio
frequency pulse ($B_{ex}(t)$):

$$\theta = \int_0^{T_{ex}} B_{ex}(t)dt.$$ 

Upon excitation, the spin system gradually returns to thermal equilib-
rium, thereby emitting electromagnetic energy. The continued preces-
sion of the net magnetization creates a time varying magnetic flux at the
Larmor frequency. This electromagnetic field can be sensed by nearby
detector coils, based on Faraday induction.

Based on the principle of reciprocity, a given magnetization ($M$) and coil
sensitivity ($C$) yields the induced signal ($S$) over a volume ($V$) spanning
positions $r$.

$$S(t) = -\frac{\partial}{\partial t} \int_V M(r, t)C(r)d^3r.$$ 

¹The actual energy difference is very small for typical $B_0$ values in MRI. At the
highest clinically used field strength of $7 \text{T} \Delta E = 12.32 \times 10^{-7} \text{eV for hydrogen nuclei}$.
The ionization energy of hydrogen in its ground state is $13.7 \text{eV}$ which is about 11
million times higher and explains why RF pulses in MRI are non-ionizing.
Relaxation coarsely involves two mechanisms: the decay of $M_{xy}$ and the recovery of $M_z$ to $M_0$. These can be modeled as first order processes with time constants $T_1$ for longitudinal recovery and $T_2$ for transverse relaxation. The received signal contains an abundance of information about the magnetized sample as several mechanisms influence this transient behavior, including the microscopic environment of the nuclei and neighboring molecules, the chemical content, or motion. Together with the actual proton (spin-) density (PD), $T_1$ and $T_2^2$ relaxation times make up the three most instrumental and often used intrinsic contrast mechanisms used in MRI.

The phenomenon of nuclear magnetic resonance (NMR) was first described by Rabi in 1938 [2] in an experiment performed to measure the nuclear magnetic moments. Few years later, in 1946, Bloch and Purcell independently observed that proper sample excitation induces spin coherence and that during relaxation an NMR signal can be captured [3], [4]. This laid the foundation for NMR spectroscopy widely used to detail the chemical constituents and electronic structure of a molecule and its individual functional groups. Still, NMR alone does not give spatial information, needed for image formation.

Therefore, in MRI, a magnetic gradient field is used to separate the bulk sample into individual voxels. A changing magnetic field ($G_{x,y,z}(t)$) superimposed on $B_0$ renders the resonance frequency of the net magnetization of atomic nuclei within the human body both position ($r$) and time ($t$) dependent, according to:

$$B(r, t) = B_0 + G_{x,y,z}(t) \cdot r$$

While gradients are active, spins that experience a higher field precess more rapidly than those at lower field, resulting in spatial frequency encoding. Similarly, when gradients are turned off, the precession frequency reverts to the initial $\omega_0$. Net magnetization that undergoes faster precession during gradient action gains a phase proportional to

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$^2$The coherence of spins immediately after RF excitation is gradually lost due to field inhomogeneities and spin-spin interactions of protons. They dephase, reducing the apparent net magnetization. The relaxation time constant for this decay is referred to as $T_2^*$. The dephasing due to field inhomogeneities can be reversed by the application of a 180° pulse, causing a spin echo and partial signal recovery.
the duration and strength of the local field seen \( \Phi(t) = \int_0^t \gamma G(t) r \, dt \) and encodes positional information in the recorded signal.

This principle of gradient encoding was first introduced by Lauterbur [5] and Mansfield [6] in 1973 who received the Nobel prize for medicine for their discovery in 2003. Erwin Hahn discovered the spin echo in 1950, and Herman Carr described and demonstrated techniques using gradients in magnetic fields for imaging in his 1952 PhD thesis [7]. Lauterbur used the backprojection algorithm prior to Richard Ernst introducing phase and frequency encoding and the Fourier transform for image calculation, in 1975 [8]. Later, in 1983, Ljunggren [9] and Tweig [10] introduced the k-space notation most widely used today.

### 1.2 State-of-the-art MRI system hardware

Over the last decades, and since the introduction of the first commercial scanners in the late 1970’s, there was a constant strive to achieve higher resolution and better sensitivity, along with reduced scan time and cost.

Today, a typical MRI scanner consists of a strong superconducting magnet, cooled with liquid helium. Cylindrical in shape it has a bore to fit the size of the sample. For human scanners 55 cm to 60 cm has long been a standard, vendors now produce machines with 70 cm bore size to alleviate claustrophobia and ease scanning of heavier patients. These magnets create the static magnetic field (measured in Tesla, T) that is needed for net magnetization to build-up. The maximum available net magnetization, whose manipulation eventually enables a signal to be received, is directly proportional to the strength of this background field. Therefore, the most straightforward option to improve the SNR is to increase the field strength. Today’s clinically installed systems have a \( B_0 \) of 0.5 T-7 T. That is around 60’000 times (3 T) the magnetic field surrounding the Earth and gives rise to Larmor frequencies in the range of approximately 21 MHz to 300 MHz. Recent ultra-high field scanners for humans reach field-strengths of 9.4 T, with the strongest research machine even aiming at 11.75 T (in the INUMAC/ISEULT project). It
is generally considered that humans\(^3\) can tolerate short term exposure to strong static magnetic fields without harm.

Inside the magnet, three additional sets of coils are installed. These \textit{gradient coils} are used to create switchable, position dependent field gradients for spatial encoding. Field gradient strengths range up to 30-50 mT m\(^{-1}\) and switching speeds are in the audio-frequency range to bandwidths of about 50 kHz. Gradient switching requires powerful amplifiers, capable of sustaining large inductive loading and reactive powers up to mega Watts. This is one factor limiting the achievable bandwidth. Recently presented strong research gradients reach slew rates up to 1200 mT m\(^{-1}\) ms\(^{-1}\) \cite{12} and enable the imaging of tissue with ultra-short \(T_2^*\). However, when humans are exposed to magnetic fields switching at very high-rates (\(\partial B/\partial t\)), agonizing peripheral nerve stimulation (PNS) may occur \cite{13}, \cite{14}, in effect limiting ultimate imaging speeds.

RF coils create the transverse electromagnetic field needed for the manipulation of the net magnetization of a sample. Low and high field (not ultra-high field, \(>4 \text{T}\)) systems typically have volume coils directly integrated into the bore. These body coils can be used for both excitation and reception, e.g. for whole-body scans. For excitation, RF amplifiers produce peak powers in the kW range (up to 30 kW) resulting in \(B_1\) fields on the order of 5-50 \(\mu\text{T}\). At the same time, the received signal strength in a receive coil is in the \(\mu\text{V}\) regime. This high dynamic range of involved signals is characteristic for MRI. Even though body coils can be used for reception, better sensitivity can be reached by using localized coils that exist for various anatomies (e.g. knee, head, cardiac, etc.).

A widely adopted strategy to boost sensitivity is array detection \cite{15}. It permits a drastic reduction in scan time with what has been established as parallel imaging \cite{16}–\cite{19}. By closely surrounding the anatomy of interest with a large number of local, decoupled receiver coils and knowledge of their sensitivities, the redundancy in information seen by each of the coils can be traded for speed. Along this strategy, channel counts of MRI receive systems have gradually increased. Scanners with

\(^3\)Apparently in contrast to the common fruit flies: researchers have reported genotoxic effects leading to mutations in the hair on their wings \cite{11}.
32-channel are standard and experimental coils up to 128 channels exist [20], [21]. These coil designs are mostly rigid and built to fit a large population of patients.

Traditionally, every receive channel of an imaging array coil is pre-amplified close to the coil and individually connected by a long coaxial cable from inside the MRI scanner Faraday cage, that prevents spurious signal pickup, to a spectrometer in the technical room, away from the harsh in-bore environment that could deteriorate acquisition. Early pre-amplification is vital to achieve acceptable SNR. MRI spectrometers employ various topologies, most receivers use the superheterodyne principle where the input signal is converted down to a low frequency of some MHz [22] prior to demodulation and digitization. The signal decays exponentially, which is why MR receivers need to provide a large dynamic range [23], [24] - to "listen" to the spin decay as long as possible.

1.3 Current challenges and trends

Despite MRI scanners being the extensively used, and complex diagnostic machines we know today, they suffer from numerous problems.

Moving to higher fields has the advantage of better SNR, yet some side effects limit broad adaptation of ultra-high field scanners. While SNR scales linearly with $B_0$, the deposited energy or specific absorption rate (SAR) increases quadratically with $B_0$. High frequency scanner operation leads to higher RF losses in the subject and shorter wavelengths. With the wavelength reaching body dimensions, the quasi-static field assumption is void. Wave effects occur and coil loading becomes non-negligible, both lead to RF field non-uniformity and potentially local RF heating caused by uncontrollable transmit fields. These pose a safety risk to patients. Especially at ultra-high field, where wave effects are becoming more pronounced, multi-transmit [25], [26], a similar concept to parallel reception, enjoys increasing popularity. It improves illumination and minimizes RF-energy deposition in tissues. Besides, susceptibility mismatch within the sample (e.g. air cavities
or tissue boundaries) give rise to local field inhomogeneities and accompanying faster signal decay. Susceptibility artifacts scale with field strength. To counteract field inhomogeneities, additional shim coils are installed. These produce higher order spatially varying fields and to a certain degree allow homogenization of $B_0$ in a sample. However, static shimming can not compensate for spurious fields of higher order or inhomogeneities over a large volume. Likewise, for high frequency operation, required by the employment of higher static magnetic fields, receiver electronics need to be improved to sustain low noise performance and the clocking of the receiver system can start to influence the maximum dynamic range achievable.

Switching gradients and shims induce eddy currents, counteracting the actual field, which hamper linearity. Equally so does coupling between gradient coils. In present day scanners, actively shielded gradient coils [27] cancel distant magnetic effects and reduce eddy currents induced in shims, the cryostat, the main coil and the housing. Although strong gradients are available and advanced calibration procedures are used, the achieved field evolution is not perfectly implemented, confounding advanced imaging. Further, only a small volume with linear fields is typically achieved. Factors continuing to limit fidelity of dynamic fields include scanner mechanical resonances, temperature dependence, and finite linearity and stability of amplifiers. The use of pre-emphasis [28] effectively counters some of these inaccuracies. However, residual errors remain. Field evolution can also be flawed by hardware deterioration, programming glitches or by physiological effects such as patient breathing, blood flow or peristaltic movements. With limited gradient fidelity, images are impaired by blurring, distortion artifacts and signal voids. Also, gradient switching produces acoustic noise, providing an unwanted stimulus and interference during fMRI studies [29] and strongly impairs patient comfort [30].

Today, the inherent potential of array detection has not nearly been exhausted and the demand for realizing ever larger channel counts persists [31]. However, increasing the number of receive coils is an engineering challenge limiting the full exploitation of parallel imaging methods. A fundamental drawback of increasing the coil numbers is the required cabling. Electromagnetic interference between the long coaxial
cables and the receivers in the equipment room can reach a serious enough level to actually deteriorate image quality [32]. Coupling among common-mode loops and the RF system can cause large currents to flow, potentially threatening patient safety, e.g. by RF burns [33]. In addition to electrical problems, the mechanical handling of these cables cumbers coil engineering. For highly sensitive arrays, coils are made smaller and need to accommodate pre-amplifier electronics in high density. These efforts get limited by stiff and bulky cables.

On the other hand, current rigid detector arrays are insufficient in terms of ergonomics and deviate from the paradigm of closely fitting anatomies for maximum sensitivity. Mechanical rigidity compromises sensitivity because it prevents adjustment to individual size and shapes of target anatomies. It prevents changes in posture such as flexion of joints and is uncomfortable for patients.

As an effective means to circumvent many of the existing hardware imperfections relating to field fidelity, observing the field evolution using additional sensors has recently been met with a great response. Field monitoring using NMR field probes measures the actual k-space trajectories, capturing the net effect of all perturbations without knowledge of the exact sources of errors [34], [35]. It allows retrospective image correction [36], [37] and has been used to monitor transmit RF [38] for better excitation field control at ultra-high fields. It was shown that fields can be stabilized using field sensors and shim updates in a feedback loop [39]. Based on NMR, field sensors require receive channels of basically the same nature as detector arrays for the acquisition of image raw data. At the same time NMR sensor signals deviate from traditional mere image acquisition by operating on different nuclei, at different frequencies and bandwidths and with different timing, requiring a versatile acquisition system.

With functional magnetic resonance imaging (fMRI) on the rise, fast imaging sequences enjoy increasing popularity. These rely on demanding gradient schemes where knowledge of the true k-space trajectory, enabled by field sensing, is crucial. Unfortunately, even the knowledge of the actual k-space trajectory is insufficient as several irreversible perturbations prevail. Signal dropouts due to dephasing, the disruption
of steady state conditions in certain sequences, or nonrecoverable signal alterations arising from subject motion, can not be corrected retrospectively. Because gradient slew rates are PNS-limited, fast sequences demand long readouts (e.g. spirals, EPI) and suffer from limited signal lifetime. Long readouts are needed for high-resolution scans and are especially prone to subject motion. Motion causes misinterpretation of acquired signals and wrong k-space coverage and is considerably one of the big challenges of MRI. Various kinds of sensors to track rigid body motion, such as the movement of the head, exist. These include cameras [37], [40]--[42], active markers [43], [44] or NMR field probes [45], [46]. Measurement of motion parameters in real-time permits continued adaptation of the scan sequence to the current patient posture. For this to work, it is critical that calculations for geometry updates are performed with minimal delay.

Clearly, many remarkable technical advancements to tackle present-day MR-system detriments were suggested. However, many of them introduce challenges themselves. Interestingly, these can be reduced to a few problems that, if solved, contribute to hugely improve MR scanner performance.

Just like array coil detection is limited by cables, presented auxiliary techniques and additional sensors exacerbate the cabling problem, associated safety, and signal integrity concerns. Removal of the large cable bundles is definitely beneficial.

Yet, cabling is not the sole challenge. For some sensors it is essential to be captured in absolute synchronicity to the image data since even minor relative delays will impair reconstructed images, as is the case e.g. with concurrent field monitoring. Others deliver information that needs to be processed and be available with lowest latency, to effectively improve system performance. This is the case for example if sensors are used to update the scanner operation in real-time as in prospective motion correction [45], [46] or background field stabilization [39].

Sensors have diverse interfaces and deliver heterogeneous data that requires different processing than conventional raw image data. A conceivable problem with the many imaging channels and sensors added, is the increased amount of data that needs to be handled, and stored.
Data pre-processing and compression are becoming ever more important. Today, first-level data processing and fusion are off-line processes, limiting e.g. real-time interventional imaging.

Contemporary MRI receive systems are not ready for the diverse multi-channel output of modern MR experiments. To overcome this and to grant the seamless combination of imaging data with auxiliary sensor data while at the same time allowing real-time processing and actuation, a new kind of MRI receiver system is necessary.

Likewise, due to lack of appropriate receivers, technical advances to improve patient comfort, such as the availability of flexible and stretchable detector coils [47], are limited in application.

1.4 Aim, contributions and outline

The fundamental goal of this thesis is to tackle challenges mentioned above, addressing MRI sensitivity, processing and patient comfort by improving three distinct shortcomings of current MRI hardware:

To greatly facilitate the implementation of detector arrays and of auxiliary instrumentation, such as magnetic field cameras, large cable bundles used in state-of-the art imaging and monitoring systems need to be removed. For this, an optically connected high performance MRI receiver is developed and experience with in-bore operation of electronics is gained. Chapter 2 describes a discrete multi-channel in-bore receiver system. These receivers are capable of being operated within the scanner bore while at full scanner operation without interfering the imaging process. The key challenge is to meet the high requirements in terms of linearity and SNR in this harsh electromagnetic environment, including static magnetic field of multiple Tesla, audio frequency fields in the range of tens of millitesla, and RF fields up to tens of kW.

To address the lack of a truly scalable receive system, capable of processing imaging and auxiliary sensor data in real-time, a flexible MRI receiver and processing system is built. Chapter 3 showcases this novel out-field hardware platform. It can be synchronized to the MRI scanner
and allows synchronous acquisition of data from multiple in-bore modules acquiring heterogeneous types of data (imaging, monitoring, and vital signs) to form a multi-channel receiver with the ability of real-time data processing for advanced imaging experiments. The implemented processing pipeline allows easy to adapt signal treatment and data flow, including programmable filtering, the synchronous calculation of field evolution from NMR signals, that also involves non-linear operations such as phase extraction, or real-time coil compression.

Ultra-high fidelity clocking for phase coherence, as well as low latency triggering are key requirements for successful MR acquisition. Chapter 4 discusses the importance of clocking and frequency translation in receiver systems and that scanner operation can degrade their performance. It is outlined that commonly used oscillator technologies are influenced by gradient operation, affecting received signal quality. A new method to mitigate gradient influence on oscillators and clock source imperfections is described and validated.

To improve patient comfort and coil sensitivity, we lay new grounds in MRI acquisition with an integrated and wearable MRI receiver system. With the knowledge gained from operating receive electronics in-bore and the possibility for custom integration, the in-bore receiver approach is taken one step further. By integration into a complementary metal-oxide semiconductors (CMOS) chip, large parts of the receive electronics could be miniaturized and operated directly on the receive coil. Chapter 5 describes successful in-vivo imaging with direct on-coil signal digitization on a flexible four-channel wrist array. This seminal step was accomplished within the NanoTera WearableMRI project, a collaboration project of three institutes of ETH Zürich.

Finally, Chapter 6 gives a summary of what has been accomplished, followed by some concluding remarks. Thoughts for potential improvements and visions for future MRI acquisition are outlined.
Chapter 2

In-bore Receiver for MRI

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CHAPTER 2. IN-BORE RECEIVER

2.1 Introduction

Since the advent of array detection [15], [48] and parallel imaging [16], [17], [19], [49], [50] MRI has seen a steady increase in the number of coil elements and receive channels deployed for signal acquisition. To maximize the sensitivity and speed of the technique, anatomies of interest are surrounded by dense, closely coupled receiver arrays. Current MRI setups typically use multiple tens of receiver elements. Systems with up to 128 channels have been described in the literature [20], [21], [31].

While modern MRI receive systems are based on scalable and digital architectures [51]–[53], scaling the number of receive coils is limited. In efforts to deploy ever more coil elements, one important obstacle is cabling. Traditionally, each coil has an individual, long coaxial-cable connection to a receiver placed away from the harsh electromagnetic (EM) environment in the magnet bore. With increasing channel numbers, management of these cables has become ever more demanding. Besides handling issues, SNR optimization and safety concerns over coupling during high-power RF transmission require escalating sophistication of cable routing and signal trapping. This challenge is made even steeper by recent efforts to monitor RF transmission [54] and magnetic field evolution [55], [56] during MRI, which likewise require RF detection with large channel counts.

As a radical departure from cables it has been proposed to transmit MR signals by wireless radio frequency links. Various approaches have been reported, including analog transmission by inductively coupled RF coil arrays [57], single-frequency multiple-input–multiple-output (MIMO) microwave links with direct frequency up-conversion [58] or systems based on digital protocols that transmit digitized data streams [59], [60]. However, as yet the wireless approach has also proven challenging in several respects, particularly in terms of SNR, data throughput, clock synchronization, and power requirements [59], [61], [62].

In these respects, optical connections offer an attractive alternative that also overcomes bulky cable trees and coupling issues while offering more controlled signal transmission and intrinsic link capacity. Several
methods of signal transmission via optical fibers have been reported [32], [63]. Most of them perform direct transmission of analog coil signal, a technique mainly limited by dynamic range, linearity, and noise figure of optical components.

These problems are best addressed by early digitization and digital transmission, which is practically immune to limits of linearity and analog SNR. In the digital domain it is also easier to exploit total link capacity with the help of digital signal processing. The data transmission capacity of a digital optical connection with driver electronics reaches several Gbps such that multiple receive channels can readily be transmitted over a single fiber.

To reap these benefits, RF detection and digitization, digital processing, and optical conversion must be performed close to the signal sources and thus in the magnet bore. Detection and digitization are particularly demanding in MRI, requiring exceptionally high dynamic range, linearity, and phase stability. Meeting these requirements with in-bore electronics is challenging because of strong static and audio frequency magnetic fields as well as high-power RF transmission, which form a particularly harsh EM environment. RF pulses can induce heating and cause signal interference or even complete destruction of electronic devices. Under switched gradients, an electromotive force (EMF) is induced in any conductive loop, which can alter supply voltages and actual signals or produce bit errors in the digital domain. The same is true for induced forces and mechanical vibrations in the audio-frequency range. Relevant materials may even have field-dependent mechanical and electrical properties. Examples include the change in Young’s modulus of the suspension of crystal oscillators that alters the fundamental frequency [64], and high-electron-mobility transistor (HEMT) devices such as, e.g., low-noise Gallium-Arsenide (GaAs) amplifiers whose gain is orientation-dependent due to the Hall effect [65]. The presence of electronics may also perturb the imaging process per se. Static field perturbation by magnetized components as well as supply and induced currents may alter resonance conditions and spin dynamics, which will cause image artifacts.
In this work we address all these challenges and present a 16-channel in-bore MRI receiver with direct digitization and optical data transmission. Performance metrics were evaluated on the bench and in 3 T and 7 T scanners. Finally it was applied for imaging and field monitoring.
2.2 Methods

2.2.1 Design considerations for in-bore electronics

Main magnetic fields of several Tesla, superimposed gradients switching at audio frequencies and RF signals with powers in the kW range demand for careful design in terms of signal integrity, power supply and printed circuit board (PCB) design. In the following, we sketch the fundamental mechanisms of distortions faced and subsequently outline how to tackle these in a design.

Mechanisms and countermeasures

In the strong background field, minimal amounts of ferromagnetic materials contained in electronic components can generate mechanical forces or distort the field homogeneity. Functioning of semiconductor devices can be dependent on their orientation, up to the point of disrupted current flow [66]. Off-the-shelf component packages often contain magnetic compounds; a typical example includes the nickel under plating used as a diffusion barrier at the leads and for improved wear resistance. In a design, non-magnetic and small components should be used whenever possible. Correct operation, also with regard to orientation, can be assessed in the scanner for every unit individually.

RF pulses for spin excitation emitted by the scanner can strongly degrade or even interrupt [62] the working of electronic devices placed in the field, e.g. by saturation of input amplifier stages or arcing of capacitors with limited voltage rating. However, in terms of RF and despite many additional design demands, in-bore or in-magnet room digitization has relaxed input filtering requirements compared to receivers outside a Faraday room, e.g. telecommunication systems, as one can assume that no strong in-band blocking signals are present.

Effective RF shielding to reduce EMI is necessary to protect electronics from the scanner and vice-versa. Yet, gradient induced eddy currents on RF shields may have adverse effects on the encoding field. Therefore, RF shields can be fragmented and re-connected using capacitors.
This reduces the direct current (DC) area while maintaining a low RF impedance for effective shielding.

Along these lines, gradients are the biggest hurdle faced when operating electronics in-bore. Faraday’s law dictates that an EMF is generated in any conductive loop that encloses a time-varying magnetic field. That is, with modern gradients reaching up to $1200 \text{ mT m}^{-1} \text{ ms}^{-1}$ [67], MRI’s high dynamic range signals [24] running on PCB traces can be modulated or distorted by induced voltages, leading to SNR loss, bit-errors or even complete failure of the circuit. With multi-layered PCBs complex current paths might lead to high local current density stress, reducing circuit reliability and life time e.g. due to excessive heat or electromigration [68]. These effects strongly depend on the actual design and layout.

A general approach to tackle gradient induction is to keep PCB attack area as small as possible and to use differential signaling as gradients predominantly act on the common mode. However, high-frequency differential transmission lines require a reference plane for uninterrupted line impedance. These reference planes, often designed as full copper pours on the layer stack, are target for eddy current induction. To mitigate arising eddy currents, a strict star-structured layout to avoid current loops must be employed, reducing area of planes to wide traces and adding slots to heat dissipation planes while taking care not to disrupt any return current paths. Moreover, ultra-low impedance power supply decoupling reduces induced supply ripple. This requires distinct capacitor variations over a large capacitance range potentially occupying significant board space. Power efficient voltage level translation with switched DC/DC converters finds limited application in-bore either because large inductors are needed due to transistor-gate capacitance limited switching speed or harmonic interference of the receive signal with leakage signal emitted by these converters. For this reason, high power supply rejection ratio (PSRR) linear regulators are used to counteract gradient induced power supply ripple and to replace typically used ferrite beads that saturate and lose their efficiency in-bore. However, low-noise supply design comes at the cost of lower energy efficiency and more heat generation.
Validation

To confirm MR compatibility of employed components and materials and to verify PCB layouts the following tests were performed:

1. Magneto-static influence and MR signal interference was assessed by imaging experiments. Components were placed at distances of 0 cm to 5 cm from a phantom and imaged using the scanner spectrometer. Image artifacts such as signal voids, ghosts, distortions or increased noise level were assessed to verify MR compatibility.

2. Correct operation of individual units under strong static and dynamic fields (3 T and 7 T scanner) was confirmed by actual in-bore validation while sine-wave tones and echo-planar imaging (EPI) sequences of maximum gradient slew rate were played out. In addition, bench gradient simulations were performed. Changing magnetic fields (up to 100's of T/s) in the audio frequency range (up to 10 kHz) were locally induced using hand-wound multi-layer solenoid coils of 0.5 cm and 1 cm diameter (124 windings) connected to a sine-wave signal generator and a 600 W audio power amplifier. Network analyzer and oscilloscope measurements as well as in-circuit digital processing analysis allowed confirmation of operability under gradients.

3. Eddy currents on conductive materials, including RF shields and PCB layout were evaluated using a custom-built eddy current test setup. It was made from a six layered coil with 288 turns, 30 mm diameter and 48 mm length. The main coil was driven by currents of different strengths (up to 4 A) and varying frequency (up to 10 kHz) to generate gradient like magnetic fields on both reference planes.
2.2.2 System design

Figure 2.1: Schematic overview of the system with in-bore receiver and acquisition system outside of the scanner room. The in-bore receiver is placed behind the coil in close vicinity to the signal source, typically around 40 cm from the iso-center. It is connected to the receive array or a set of field monitoring probes. Data are transmitted by an optical fiber to an acquisition and processing platform for further processing and storage. The platform provides a high fidelity clock through the back-channel of the optical fiber and generates trigger signals transmitted via a second optical fiber to the in-bore modules. In this way, multiple in-bore receivers can be synchronized to the scanner and among each other. The single remaining galvanic connection in-bore is the power cable to the in-bore receiver.

The acquisition system comprises of in-bore receivers connected via a bi-directional, high-speed optical link to a acquisition and processing platform outside of the scanner room, which is discussed in more detail in Chapter 3. Figure 2.1 shows a schematic overview of the system.

The receive architecture is based on direct RF digitization with band-pass sampling [69], where the band-limited input signal is sampled at a lower rate and intentional aliasing is exploited, replacing traditionally used mixing stages. Classical analog operations such as down conversion, demodulation, and filtering are entirely performed in the digital
domain, reducing electronics component counts and circuit complexity. Real-time digital signal processing is realized on a FPGA with custom cores.

2.2.3 Hardware

Overview

Figure 2.2 shows the schematic of the proposed in-bore system. The hardware comprises of: 1) a baseboard (10 cm x 12 cm) composed of an FPGA, clocking circuitry, communication lasers and two extension connectors, 2) two 8-channel analog-to-digital converter (ADC) boards of the same size as the FPGA board with differential amplifiers and digital input-output (IO) control, and 3) 16 RF gain scaling and filtering modules (4 cm x 3 cm) that are mounted on the ADC boards. Photos of corresponding PCBs are shown in Figure 2.3 and a fully assembled in-bore unit is shown in Figure 2.4a. It is housed in a custom box of 10 x 11 x 12 cm\(^3\), lined with thin conductive fabric for RF shielding (Holland Shielding, Dordrecht, The Netherlands) as shown in Figure 2.4b.

All PCBs are multi-layered and based on improved FR4 glass-reinforced epoxy laminate and chemical silver refined copper traces. To ensure signal integrity for data rates over 3 Gbps, all high-speed digital and analog trace dimensions on the layer-stack were simulated using a 2.5-dimensional (2.5D) EM field solver (TNT-MMTL) with post-production factory validation. For high-density ball grid array (BGA) fan-out on both the ADC- and FPGA- boards, filled, buried and micro-vias were used.

Gain modules

Two types of pluggable gain scaling and filtering modules were designed, for 3 T and for 7 T. The modules cover both \(^1\)H and Fluorine-19 (\(^{19}\)F) frequencies for both imaging and concurrent field monitoring applications. They have a low noise figure (NF) (0.95 dB @128 MHz, 1.1 dB @298 MHz) and a programmable gain range of \(-15\) dB to 40 dB.
Figure 2.2: Schematic overview of an in-bore receiver. It comprises of three board types: a) 16 application specific gain modules (shown here a default implementation for imaging and field monitoring having two amplifiers, bypass switches, an attenuator, and analog RF band-pass filter for band pre-selection in between); b) 2 high-speed multi-channel ADC boards for signal digitization (125 MSps 8-channel ADC) with input protection and anti-aliasing low-pass filters; c) an FPGA (Artix-7) board for signal down-conversion and high-speed serialization, a precision clock circuitry and optical modules for configuration, triggering and data streaming.
For both 3 T and 7 T applications a first stage fixed gain low-noise amplifier (LNA) (SPF5043Z, Qorvo, Greensboro, NC, USA) is followed by a second RF amplifier (TRF37A73, Texas Instruments, Dallas, TX, USA), a 3rd order Chebycheff band-pass filter with a bandwidth (BW) of 41 MHz for 3 T and of 57 MHz for 7 T, a digital step attenuator (PE4312, pSemi, San Diego, CA, USA) and a third amplifier stage, equivalent to the second. Two switches (PE4251, pSemi, San Diego, CA, USA) allow bypassing of the second and third amplifier stage for linearity or noise figure optimization at defined gain. For a gain setting where either one of the amplifiers is bypassed, the inherent trade-off between noise figure and linearity can be balanced depending on the application according to Friis’ formula for noise [70]. The modules include a transmit-receive (T/R) switch for e.g. field probe excitation with an externally fed RF signal.

**Analog-to-digital converter board**

Input signals are fed via short cables (typ. 40 cm) to the 6-layer ADC board via micro-coaxial (MCX) connectors. After a power limiting input protection diode and an input termination switch, the signals are amplified and filtered in the gain-modules which are plugged into the ADC board. Besides RF connectors, a second connector provides power supply and digital control. After amplification, signals are converted from single-ended to differential using high-linearity, dual channel broadband differential amplifiers (ADL5566, Analog Devices, Norwood, MA, USA). To reduce harmonic aliasing a fully-differential custom low-pass anti-aliasing filter (3 dB-BW = 498 MHz, 3rd order Butterworth) is placed in front of the ADC (LTM9011, Linear Technology, Milpitas, CA, USA). Iterations on PCB layout and gradient modulation tests showed that the fully differential design experiences significantly less modulation (< −40 dB) under gradients compared to single-ended signaling.

The ADC offers eight synchronously sampled input channels with an input bandwidth of 800 MHz and a nominal resolution of 14 bit with an effective number of bits (ENOB) of 11.9 bit at the desired input frequency of 128 MHz and 11.4 bit at 298 MHz. The sampling rate is programmable 5-125 MSps. It has a power consumption of 140 mW.
per channel. High-speed, low voltage differential signaling (LVDS) 16 bit serialized outputs provide digitized data to the interface connector. Two non-volatile flash-based FPGAs (Igloo nano, Microsemi, Aliso Viejo, CA, USA) enable synchronous ultra-fast digital IO control, including input protection switching and field-probe T/R switching. Communication takes place via an I2C bus and dedicated trigger lines for low-latency. Power inlet to the board is 2.5 V and 4 V and on-board supply regulation (1.5 V for the FPGAs, 1.8 V for the ADC, 3.3 V for the amplifiers and gain modules) is achieved exclusively with low-dropout regulators (LDOs) with ultra-high PSRR and low output noise (LT3080, LT1965, Linear Technology, TS7A4700 Texas Instruments).
Figure 2.3: Photograph of hardware components: a) Gain module with three amplifier stages, band-pass filter and digitally programmable attenuator. b) 8-channel ADC board. Signals are received through MCX connectors. After input protection and gain-modules, signals are fed to a broadband linear differential amplifier, which converts the signals from single-ended to differential, followed by a 500 MHz anti-aliasing filter and the 8-channel ADC. c) FPGA board with clock generator chip and FPGA (Artix-7), high-speed bi-directional optical module, optical trigger input, two expansion connectors (front- and backside), and opto-galvanically isolated IOs.
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FPGA board

The 10-layer FPGA baseboard connects extension boards (ADC boards) via two connectors with a dedicated differential clock output, 23 LVDS lanes, 12 GPIOs, I2C, JTAG, and power (2.5 V, 4.0 V, adjustable power) to allow versatile use. Power supply is decoupled with a wide capacitor range amounting to a total of 3 mF to provide a low supply impedance over a wide frequency range to avoid gradient influence. It hosts an FPGA (Artix-7, Xilinx, San Jose, CA, USA) to process the large amount of data arising when sampling 16 channels at 125 MSps (16 x 16 bit x 125 MSps = 32 Gbps data). Multiple in-bore receivers can easily reach raw data rates of hundreds of Gbps, an amount of data that can hardly be transmitted or stored in real-time. Therefore, digital down-conversion is performed before signal transmission (see digital processing section below). Down-converted and decimated data of all channels are transmitted as a serialized data stream over two lanes.

The total link data rate of 5 Gbps (2 x 3.125 Gbps with 8b10b coding) sets an upper limit to the BW per channel depending on word size after decimation (W) and number of acquired channels (N):

\[ BW = \frac{5 \text{ Gbit}}{W/N}, \text{yielding } 4.89 \text{ MHz for full duty cycle acquisition of 16 channels with } 2 \times 32\text{-bit words (I/Q data).} \]

The employed optical modules (P1TX6B, Inneos, Pleasanton, CA, USA) are based on wavelength multiplexing (778 nm, 800 nm, 825 nm, 850 nm, 911 nm) of five optical channels on a single fiber. Each module incorporates four high-speed (3.125 Gbps) unidirectional lanes and a bidirectional channel (1.25 Gbps). Data are transmitted via two of the high-speed channels and the bi-directional channel is used for reference clock transmission and communication.

A high-precision phased locked loop (PLL) (Si5345, Silicon Labs, Austin, TX, USA) with zero-delay capability for optimal phase coherence and easy frequency planning is used as clock generator. For stand-alone operation, a low-drift temperature-compensated crystal oscillator (TCXO) (TB512-050.0M, Connor-Winfield, Aurora, IL, USA) can be enabled. When under-sampling is employed, the optimal sampling frequency \( f_s \) depends on the input frequency. The PLL’s flexible frequency
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generation allows careful selection of the sampling frequency to reduce unintended aliasing. For 7 T operation \( f_n = 298 \,\text{MHz} \), the ADC's sampling rate is selected to be 125 MSps with the alias located at 48 MHz. The sampling rate for 3 T operation is 117.1875 MSps achieved by integer division in the PLL, placing the alias at \( \approx 10.5 \,\text{MHz} \).

The baseboard hosts 2 Gbit DRAM and 128 Mbit flash-memory to store the logic bit-stream and application software. Three trigger outputs with switchable impedance and two optically coupled trigger inputs are used to interface to other electronic devices in or out of bore. For synchronous triggering and sequencing of several modules, the module has an optical trigger input (HFBR, Avago, San Jose, CA, USA).
Figure 2.4: A fully assembled in-bore receiver. a) A photograph of a fully assembled in-bore receiver with all modules mounted, including (a) gain modules, (b) ADC boards, and (c) FPGA board. b) In-bore receiver housed in a compact 3D printed enclosure.
2.2. METHODS

2.2.4 Digital subsystem

The digital subsystem is designed similar to a micro-controller system. The processor controls the cores registers via a system bus. In this way, the functionality of cores can be controlled via software. In this design, the MicroBlaze processor (Xilinx, San Jose, CA, USA) and AXI4-MM bus (ARM, Cambridge, UK) from the vendor’s standard library were used. After power up, the processor configures the PLL, the ADC and amplification modules and initializes the software drivers of the cores. Subsequently, it enters a software routine for receiving, executing and transmitting commands from the acquisition system operating outside the scanner room (Chapter 3). For this purpose, a custom protocol with fixed length commands and a high-speed UART core were implemented. An overview of the FPGA architecture is shown in Figure 2.5a.

Two cores were custom designed, a digital-down-conversion (DDC) core and the high-speed data link (HSDL) core for data transmission via the optical link (Figure 2.5a, blue and green). The DDC consists of three distinct parts, a trigger generator for sequencing and synchronization with external devices, a register interface through which the operation of the core is controlled, and the signal processing chain. The signal processing chain is shown in Figure 2.5b. A sampling logic deserializes the incoming data streams from the ADC. It has automatic deskewing logic implemented to compensate for temperature-dependent delay changes and different trace lengths on the PCB and in the FPGA. Since the ADCs deliver data on four different clock domains (two ADCs with two serialized lanes each), resynchronization to the system clock (156.25 MHz) is performed using first-in-first-out (FIFO) buffers. The demodulation frequency of the receiver is set via a software register in form of a phase increment. Its constant accumulation generates a slope, which is input to a CORDIC in rotation mode [72]–[74]. The CORDIC generates two sinusoids with 90° phase shift which are used for quadrature demodulation [75] of the re-sampled input signal using two 25 x 18-bit block multipliers. The rounded output is filtered using a 4th-order cascaded-integrator-comb (CIC) filter [76] with adjustable decimation factor. The CIC introduces a decimation factor dependent gain that is corrected for by adjusting the most significant bit (MSB) of
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Figure 2.5: Digital Schematic. a) Architecture of the system on chip on the FPGA fabric. It is based on the soft-core processor (MicroBlaze) (orange) and peripheral cores connected with a standard bus (AXI4-Lite). A C-application running on the processor controls the hardware logic, ensures correct configuration of cores and communication. Data is sampled from the ADC high-speed interface by a custom DDC and pulse generation core (light blue). After signal-processing data is transmitted via a high-speed serializer core (light green) to the gigabit optical transmit link. b) Detailed view of the digital down-conversion core. The signal path is marked with arrows with the corresponding channel and bit-width where applicable. A CORDIC circuit is used to generate the quadrature demodulation signal. Demodulated signals are passed to a CIC filter with programmable decimation and automatic gain compensation followed by a FIR filter with programmable coefficients and decimation factor. All data types can be stored in the output buffer to allow for multi-rate signal acquisition [71]. The sequence generator is used to start the acquisitions and to time critical triggering for e.g. field-probe excitation.
the 32 bit output word. Further filtering and decimation is performed using a finite impulse response (FIR) filter with software programmable coefficients and decimation factor. The filter is implemented as direct form FIR filter, which utilizes a single multiply-accumulate (MAC) per channel. Decimation is implemented by calculating only necessary convolutions. This enables optimal use of the available hardware resources with a maximal filter length $N_{\text{max}}$:

$$N_{\text{max}} = \frac{f_{\text{sys}}}{f_s} \cdot D_{\text{CIC}} \cdot D_{\text{FIR}}$$ (2.1)

where $f_s$ is the sampling frequency of the ADC, $f_{\text{sys}}$ the system clock the FIR filter is operating on, $D_{\text{CIC}}$ the decimation factor of the CIC filter, and $D_{\text{FIR}}$ the decimation factor of the FIR filter. The outputs of the CIC, FIR filters, and the re-synchronization buffer filled with raw samples are connected to a buffer module. Control logic is implemented to select channels and samples of the three data types (as needed e.g. in [71]). Triggered by the rising edge of the acquisition trigger a number of reject samples are skipped and acquisition samples fed into the respective buffer. If more than one data type is selected, the data types with a higher sample rate are transmitted first, i.e. raw data first, then CIC, and FIR data. On each acquisition trigger rising edge the filters are reset in order to have a constant group delay throughout all readouts. Data is passed through a AXI4-Stream bus (ARM, Cambridge, UK) to the HSDL. It embeds the Xilinx Aurora core with a register interface added for control of the core state, error flags, re-initialization, etc. The core uses two multi-gigabit transceivers (GTP) operating at 3.125 Gbps, which pack the data in an 8b10b DC-balanced protocol for high-speed data transmission.
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2.3 Clocking and synchronization

Jitter and SNR

A practical approximation for the maximum achievable SNR of an \( N \)-bit ADC can be modeled by three factors \cite{77}: the effective input noise \( V_{\text{RMS}} \), the average dynamic nonlinearity (DNL) \( \epsilon \) and the total jitter including aperture \( t_{\text{ja}} \) and clock \( t_{\text{jc}} \) jitter:

\[
SNR = -20 \log_{10} \sqrt{(2\pi f_{\text{in}} t_{\text{RMS}})^2 + \frac{2}{3} \left( \frac{1 + \epsilon}{2^N} \right)^2 + \left( \frac{2\sqrt{2} \cdot V_{\text{RMS}}}{2^N} \right)^2},
\]

(2.2)

where \( t_{\text{RMS}} = \sqrt{t_{\text{jc}}^2 + t_{\text{ja}}^2} \) is the total jitter and \( f_{\text{in}} \) the analog input frequency of a full-scale sine-wave. \( N, \epsilon, V_{\text{RMS}} \) and \( t_{\text{ja}} \) are defined by the integrated circuitry and have seen a drastic increase in performance over the past years. Large input bandwidths of available high-performance ADCs render the phase noise of the sampling clock the dominant source of dynamic range degradation.

For the case with quantization noise only and with ideal sampling clock \((t_{\text{RMS}} = 0, \epsilon = 0, V_{\text{RMS}} = 0)\), the model reduces to the familiar: \( SNR = 6.02N + 1.76\text{dB} \) \cite{77} and since ADC parameters are fixed and input noise given, SNR in terms of input frequency and total jitter can be written as:

\[
SNR_{\text{max}} = -20 \log_{10}(2\pi f_{\text{in}} \cdot t_{\text{RMS}})
\]

(2.3)

Sampling clock phase noise has a spectral density \( T_j(f) \). Assuming a modulation of the clock signal by timing jitter around the main frequency \( (f_0) \), we can define the root mean square (RMS) clock jitter \( (t_{\text{RMS}}) \) as:

\[
t_{\text{RMS}} = \frac{1}{2\pi f_0} \sqrt{\int_{f_{\text{BW}} / 2}^{f_{\text{BW}} / 2} 2T_j(f) df},
\]

(2.4)

Please note that only jitter components in the final signal bandwidth \( (f_{\text{BW}}) \), bound by the total observation time \( (T_{\text{meas}}) \), contribute to signal degeneration.
A practical example: to achieve a receiver SNR of 74 dB with a 128 MHz input frequency, a maximum jitter of 248 fs is tolerable, while at 300 MHz only 105 fs are allowed. Consequently, extreme care has to be taken to keep the sampling clock’s phase noise minimal. The same applies for the local oscillator phase noise of mixer stages traditionally used in heterodyne receivers.

**Clock generation, transmission and cleaning**

MRI is a phase sensitive imaging modality and coherence among receive and transmit chain crucial. Due to limited long-term stability and phase noise performance of the integrated voltage-controlled oscillator (VCO), relying solely on local clock generation leads to poor long-term stability and hence undesirable image artifacts (Figure 2.6, yellow line). To ensure multiple in-bore modules are synchronously operating, an ultra-low phase noise and long term stable common reference clock is fed to the in-bore modules via the up-link of the high-speed optical link (Figure 2.6, black line).

However, the link deteriorates the phase noise of the transmitted clock signal and cannot directly be used as sampling clock with the ADC as the jitter reduces the maximum dynamic range beyond acceptable bounds (Figure 2.6, blue line, eq. 2.3). Phase noise degradation of the optical link stems from quantum noise, limited optical transmit power and associated noise in the receiving photo-diode and driving electronics. Phase noise degradation is mostly present at higher frequency offsets. Therefore, a local PLL chip cleans high frequency jitter contributions while maintaining long-term stability of the externally fed reference. Figure 2.6, red line shows the phase noise plot of the resulting clock signal.
Figure 2.6: Sampling clock phase noise. At high input frequencies, sampling clock jitter is the dominant factor defining maximum SNR. With long read-out times, long term stability of the clock is important for correct image formation. Even the best in class local oscillators have a high 1/f phase noise content, reducing required long term stability (yellow). A high precision reference clock (black) is deteriorated by the laser and photo diode driving electronics phase noise (blue), increasing the timing jitter by orders of magnitude. To counteract this deterioration, a local PLL is used to clean up the high frequency phase noise contribution of the optical link to the performance level of the local oscillator and while retaining the good long-term stability provided by the external reference, yielding good overall sampling clock performance (red).
2.3. CLOCKING AND SYNCHRONIZATION

2.3.1 Bench measurements

Dynamic range performance was assessed by acquiring a sine wave of 128 MHz and 298 MHz respectively at optimum input power and sampling frequency for following hardware configurations: 1 & 2) system equipped with the implemented gain-modules for 3 T and 7 T respectively and 3) configured without gain modules. Decimation factor was 128 for all measurements ($D_{CIC} = 32, D_{FIR} = 4$) yielding a processing gain of 21 dB and a final signal bandwidth of 915 kHz for 3 T and 976 kHz for 7 T respectively. SNR figures were calculated based on power spectral density with total noise power being integrated from 100 Hz to full bandwidth.

For spur free dynamic range (SFDR) calculation, the difference of the main signal (demodulated to DC) to the strongest frequency spur within the final signal bandwidth above 2 kHz from DC was considered. Linearity was measured by a two-tone test with 100 kHz signal spacing using two signal generators (SMA100A & SMB100, Rohde & Schwarz, Munich, Germany) and a hybrid coupler at $-50 \text{ dBm}$ total input power to the modules. Noise figure of gain-modules was measured using a noise figure meter (HP8970A, Hewlett-Packard, Palo Alto, CA, USA) and two noise sources with equivalent noise resistance (ENR) of 15 dB and 5 dB respectively.

2.3.2 Receiver influence on scanner and vice versa

To validate correct operation and performance in the scanner bore, the following tests and measurements have been performed on commercial scanners (3 T & 7 T Achieva, Philips Healthcare, Best, The Netherlands):

Scanner influence on optical link and gradient modules

Correct operation of the optical link under influence of RF pulses and gradients was tested by transmitting pseudo-random number sequences
continuously for over an hour from the in-bore receiver to the outfield station where they were compared for bit error measurements.

Components of the gain modules were individually tested on their static field and orientation dependence by comparing their performance when operated in-bore or on the bench.

**Gradients influence on receiver clock performance**

Influence of gradient operation on the received signals was tested. Combinations of gradient tones within the limits of maximum available amplitude \(40 \text{ mT m}^{-1}\) and slew rate \(200 \text{ mT m}^{-1} \text{ ms}^{-1}\) were played out on all gradient axes (separate x, y, z, and combinations thereof) while recording a reference tone close to the \(^1\text{H}\) frequency (295 MHz) to avoid NMR signals, and a power of 3 dB below full-range (signal generator: SMA100, Rohde&Schwarz, Munich, Germany). At a position of 40 cm from the iso-center, gradients are non-linear and fields have higher than nominal dB/dt, exposing the electronics to worst-case induction along the z-axis. During gradient tone play-out, the board was oriented such that the largest PCB area was exposed normal to the scanner’s x-direction.

**EMI assessment of receiver electronics**

For system RF interference assessment, images of a mineral oil phantom were recorded at 7 T using a 16-channel head coil (Nova Medical, Wilmington, MA, USA). Gradient-echo sequences with following scan parameters were employed: echo time (TE) = 1.9 ms, repetition time (TR) = 80 ms, flip angle = 60°. The images were recorded with the in-bore receiver module placed at 40 cm distance from the iso-center and compared with images without the receiver in the scanner bore. Additionally, images were made with the in-bore receiver.
2.3. CLOCKING AND SYNCHRONIZATION

Receiver influence on the static magnetic field

Receiver influence on the static magnetic field of the scanner was assessed by acquiring $B_0$-maps ($TE = 3\text{ ms}$, TE difference = 1 ms) of a spherical phantom (mineral oil) at a 7 T scanner and positioning the in-bore receiver at varying distances from the iso-center (20 cm to 3 m).

Receiver influence on the dynamic magnetic fields

Distortions of the encoding fields were evaluated using a commercial field camera with 16 $^1\text{H}$ field-probes (Skope MRT, Zurich, Switzerland) and comparing the trajectory read-out of an EPI and a spiral sequence with and without the receiver operating in-bore (40 cm from iso-center).

2.3.3 Imaging and field sensing experiments

To interface standard coil-arrays on commercial scanners, a custom-built coil interface box containing a bias-T, supply and manually switchable RF switches (PE4231, pSemi, San Diego, CA, USA) was developed. The 16-channel circuit connects malfunction and de-tune signals, as well as pre-amplifier supply of the scanner interface to the coil and couples the coil signal either to our receive system or the scanner. This enables fast receiver selection and direct comparison of acquisition performance while retaining coil safety features and scanner usage without custom patches. A 16-channel head coil (Nova Medical, Wilmington, MA, USA) and a T/R front-end with 16-field probes [56] were used.

To demonstrate in-vivo imaging capability of the receiver with synchronous acquisition at high-duty cycle and high dynamic range, magnetization prepared rapid gradient echo (MP-RAGE) and gradient-echo sequences with spoiled turbo field echo (TFE) were used. These sequences have been shown to generate signals with the highest dynamic range [24]. Brain images of healthy volunteers were acquired at 7 T.
Reconstruction was performed in Matlab (MathWorks, Natick, MA, USA) using only fast fourier transform (FFT) and a tapered cosine k-space filter. Sequence phase synchronization was achieved by recording the excitation pulse at lower gain setting and extracting its phase by a least-squares linear fit [78], [79].

To assess the ability to record field probe data both EPI and spiral sequences were monitored with the in-bore receiver and field-camera front-end with T/R box presented in [56]. Field probe signals of 16 $^{19}$F probes [35], [55] were fitted to spherical harmonics up to third order. The in-bore receiver FPGA controlled T/R switching and excitation of all field probes by means of a trigger signal to a standalone version of the direct digital synthesis (DDS) module presented in [56].
2.4 Results

2.4.1 Bench measurements

Total integrated phase noise of the sampling clock on the in-bore unit was 512 fs (equiv. 0.4 mrad$_{\text{RMS}}$ at 125 MHz). Channel cross-talk was $-48.3$ dB in the worst case for neighboring channels sharing the same differential amplifier and below $-60$ dB across all the remaining channels. Table 2.1 and 2.2 summarizes the receiver performance metrics. Measurement results confirm that the phase noise of the sampling clock becomes the dominating factor for maximum achievable dynamic range at high input frequencies. The discrepancy in dynamic range of operation with and without gain modules is attributed to the additional NF presented by the input protection and switch circuitry and the reduced available SNR of the signal generator at lower output power setting. Power consumption of a fully configured 16-channel module operating at 100% duty cycle was 22.7 W or the equivalent of 1.42 W per channel.

<table>
<thead>
<tr>
<th>Gain modules</th>
<th>3 T ($f_{\text{in}} = 128$ MHz)</th>
<th>7 T ($f_{\text{in}} = 298$ MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SNR [dB]</td>
<td>87.8</td>
<td>90.0</td>
</tr>
<tr>
<td></td>
<td>80.2</td>
<td>83.4</td>
</tr>
<tr>
<td>SFDR [dBc]</td>
<td>88.3</td>
<td>98.4</td>
</tr>
<tr>
<td></td>
<td>85.1</td>
<td>92.5</td>
</tr>
<tr>
<td>IMD3 [dB]</td>
<td>84.3</td>
<td>96.0</td>
</tr>
<tr>
<td></td>
<td>75.3</td>
<td>89.7</td>
</tr>
<tr>
<td>Gain module NF [dB]</td>
<td>0.95</td>
<td>1.1</td>
</tr>
</tbody>
</table>

*Table 2.1: Receiver Bench Performance Results*
2.4.2 Receiver influence on scanner and vice versa

Scanner influence on optical link and gain modules

Correct operation of the optical link under influence of RF pulses and gradients was tested continuously for over an hour and no bit errors could be observed.

Evaluation showed that e.g. the SPF-5122Z (Qorvo, Greensboro, NC, USA) with better noise figure has higher background field gain dependence (up to 1.5 dB) than the SPF-5043Z (Qorvo, Greensboro, NC, USA).

Gradients influence on receiver clock performance

Figure 2.7 shows the gradient induced modulation on a recorded tone during an EPI read-out with full gradient slew-rate in x-gradient direction (worst-case). The standard deviation of the residual magnitude, i.e. noise floor, changed 2.4 dB due to gradient action in the worst case, confirming good power supply decoupling and PCB layout. Gradient induced phase modulation amounted up to 35 mrad in the worst case and was below 4 mrad otherwise.
Figure 2.7: Gradients influence on in-bore sampling electronics. Modulation is strongly dependent on the orientation and positioning within the scanner bore. Under worst-case conditions (highest dB/dt in the direction of largest spanned area), the standard deviation of the residual amplitude increased by 32% or an equivalent of 2.4 dB. The signal phase modulation amounts to up to 35 mrad\textsubscript{p-p}. Bench experiments show that mostly the VCO control voltage and power supply rails are susceptible to gradients, modulating acquired phase signals.

Please note that gradient action does lead to a modulation and not to additive noise on the incoming signal. The modulation of the phase was identical for all 16 acquired channels, indicating a common phase modulation source. Bench gradient measurements confirmed that the reference clock phase can be modulated by induction. This is attributed to direct crystal microphonic modulation [64] and to the fact that even with minimized induction area some current loops cannot be avoided in an electronic circuit e.g. the single-ended routing of the VCO control voltage or the VCO supply rail.

**EMI assessment of receiver electronics**

Figure 2.8 shows phantom images recorded using the scanner spectrometer with the receiver in-bore and without it, and their corresponding log-log-images. Those serve the purpose to suppress very high values and accentuate very weak features. Even with such drastic measures no RF
Figure 2.8: Phantom images acquired using the scanner spectrometer with the receiver in-bore and without it, and their corresponding log-log-images. Additionally, an image made with the in-bore receiver was acquired for qualitative comparison. No RF artifacts, distortions or increased noise level is observed by the presence of the in-bore system.

Artifacts that typically stem from clocking or high-speed digital switching signals could be observed, indicating good shielding towards the scanner, low on-board EMI (e.g. digital to analog circuitry cross-talk) and overall good in-bore imaging performance. Additionally, an image recorded with the in-bore receiver is shown for qualitative comparison.
2.4. RESULTS

Receiver influence on the static magnetic field

Differences of the standard deviations of a reference $B_0$ map without the receiver being present and when running were calculated and showed 8.4 ppb (equiv. 2.5 Hz) at 20 cm and 3.2 ppb at 40 cm distance from the iso-center, a typical position for placing an in-bore receiver. Some of the components have a magnetic moment and influence $B_0$, e.g. the optical link modules are enclosed in a slightly paramagnetic Zamak casing (an alloy made from zink, magnesium and copper). However, measured field distortions of 10 ppb and lower are less than e.g. the $B_0$ drift seen over some minutes of scan time [80] or induced by temperature change by gradient heating [81]. For subsequent tests and imaging experiments, the receiver was kept at the distance of 40 cm from iso-center.

Receiver influence on the dynamic magnetic fields

Calculated k-space trajectory deviations from a spherical harmonic fit up to third order were below the inherent field camera measurement accuracy and scanner gradient imperfections, indicating the influence of the in-bore receiver on the encoding fields to be negligible.

2.4.3 Imaging and field sensing experiments

Images were acquired with the in-bore receiver placed at 40 cm distance from iso-center. Figure 2.9a shows 5 out of 57 slices of a longitudinal relaxation time ($T_1$)-weighted 3D-MP-RAGE image of a healthy volunteer acquired with 2 averages. Scan parameters were as follows: field of view (FOV)=220 x 220 mm$^2$, voxel size=0.7 x 0.7 x 1.4 mm$^3$, flip angle=7°, TR=6.5 ms, TE=2.9 ms. Figure 2.9b shows 5 out of 15 slices of transversal relaxation time ($T_2^*$) weighted gradient-echo image of a healthy volunteer (FOV= 210 x 210 mm$^2$, voxel size= 0.5 x 0.5 x 2.0 mm$^3$, flip angle=35°, TR=450 ms, TE=20 ms).

Figure 2.10 shows a monitored trajectory of a spiral recorded in-bore. The probe phases were fit on spherical harmonics up to the third order according to [55].
(a) High duty cycle and high dynamic range image acquisition of a $T_1$ weighted 3D-MP-RAGE sequence. Images were acquired with the in-bore receiver placed at 40 cm distance from iso-center. Every tenth slice of the stack is shown with a spacing of 14 mm (5 out of 57 slices).

(b) $T_2$ weighted gradient-echo image of a healthy volunteer. Images were acquired with the in-bore receiver placed at 40 cm distance from isocenter. Every third slice is shown with a spacing of 5.2 mm (5 out of 15 slices).

Figure 2.9: In-bore in-vivo imaging, demonstrating good acquisition performance of the in-bore receiver.
Figure 2.10: A single-shot spiral sequence monitored with 16 $^{19}$F field monitoring probes, acquired with the receiver placed in-bore. Spherical harmonics up to the third order in space are shown.
2.5 Discussion

The trend to ever more channels and heterogeneous signal sources can be addressed effectively with compact in-bore digitization. In this work, a compact sized in-bore receiver with 16 RF channels was presented. It replaces long, cumbersome and potentially dangerous cables with a single optical fiber leaving the power supply as the only remaining galvanic connection. With careful layout and component selection, it was developed to offer competitive signal quality even in the harsh EM environment inside an MRI scanner bore and without disturbing scanner operation.

The system has only weak influence on the static magnetic field $B_0$ even at ultra-high field (7 T) and when placed as close as 20 cm to the imaging volume. RF artifacts due to receiver in-bore operation could not be observed in images acquired with the scanner’s default spectrometer and in-bore receiver. The acquisitions showed sufficient quality even for the most demanding applications nowadays, i.e. for high dynamic range imaging and field monitoring.

Receiver dynamic range was found to be constrained by the clocking accuracy of the system and in special cases can be hampered by scanner gradient induction. This does not pose problems for most commonly used imaging sequences and may only have negative influence in more exotic acquisition schemes where gradients are being switched at high slew rates during read-out, e.g. spirals. For field monitoring this gradient induced phase modulation can become problematic if field probe SNR approaches receiver dynamic range. However, typical initial field-probe SNR of 40 dB [56] is well below the remaining receiver dynamic range, leaving measurements probe-SNR limited. Nevertheless, it is desirable to improve the clocking system for ultimate performance. Alternatively, the phase modulation can be removed in post-processing using a clock feedback scheme as will be detailed in Chapter 4.

A limiting factor of the in-bore receiver is the relatively high power consumption. Apart from the high frequency digital data processing requirements, the comparably inefficient LDO-only power supply is a key cause for this. LDOs provide ultra-low noise power rails and with
2.5. DISCUSSION

High PSRR contribute considerably to the gradient immunity of the analog front-end. Yet, their voltage level translation efficiency is low. Replacing them by more efficient siblings (e.g. charge pump regulators) would allow for a drastic reduction of overall power consumption and heat generation. However, more efficient alternatives generate noise and may disturb the sensitive measurements. Moreover, charge pumps are only available with small power ratings, whereas switching power supplies require large inductors. Since only air coils can be used, which leave a lot of vulnerable area, gradient modulation could impair the circuit stability. Nevertheless, power supply design remains an area to be explored in future work. In any case, the reported worst case power consumption can readily be reduced by simple measures: since MR sequences seldom reach more than 50% acquisition duty cycle, processing could be throttled during idle times as FPGA power consumption scales linearly with processing speed (CMOS technology) with a base consumption of 0.21 W per channel. Further, a full gain setting and hence maximum power requirement of the analog part (0.5 W per channel at max gain) is not needed in most cases. Although, these improvements remain to be analyzed and implemented, the presented in-bore system is an all in one receiver solution and is ready to be conveniently embedded in coil casings and field camera systems.

The scalability of the presented receiver allows the acquisition of massive amounts of channels in the ultimate vicinity of the signal source. It has the ability to handle all channels with high bandwidth and duty cycle, and enables the acquisition of a wide range of input signals. The re-configurable nature of the digital subsystem based on FPGA permits a custom design of the signal processing pipeline. This versatility empowers applications such as multi-rate acquisition used for zero-echo-time (ZTE) imaging in [71] or the simultaneous recording of multi-nuclear signals. Hardware modularity renders the system a convenient starting platform for future projects.

Placing and operating electronics in-bore requires thoughtful design and is one of the key steps towards the ultimate goal of multi-sensory, untethered and comfortably wearable MRI coils. This work sets the foundation for multi-channel in-bore digitization and gained insight enables further advances in the field of in-bore sampling.
Chapter 3

Scalable MRI Receiver and Processing System

Partially published in:


*Authors contributed equally to this work.
CHAPTER 3. MRI PROCESSING SYSTEM

3.1 Introduction

In present-day MRI scanners, arrays with up to 64 coils have become clinical standard. In favor of higher image quality and faster imaging [15]–[17], [19], [48]–[50], the trend is towards even more coils [20], [21], [82]. This is associated with increased cabling complexity and issues with signal quality and safety [33]. To solve this problem, it has been proposed to perform the acquisition and first stage data processing closer to the coil. Moving the spectrometer in-bore [83]–[86], the analog coil signals are transformed into a digital data stream and transmitted either optically [83]–[86], or by a wireless link [59].

With the increasing number of RF channels, the amount of data is growing posing challenges for throughput and data handling, which is becoming a limitation. The trend towards high acquisition duty cycles and bandwidths common for single-shot readouts [87]–[91], or ZTE imaging [71], [92] further increases the data load. For reduction of the data amount, early pre-processing and compression [93], [94] is becoming inevitable.

In recent years MRI scanners are increasingly complemented with sensors such as active markers used for catheter tracking [95], motion correction [43]–[45], [96], and field monitoring [35], [55], [56], [97]–[99] as well as sensors based on other technologies like motion tracking cameras [37], [40]–[42], systems for respiratory and cardiac triggering [100]–[104] and eye tracking [105].

Sensors not only contribute to the data and growing RF channel strain, but also deliver heterogeneous data, which need different processing than imaging data and precise synchronization with it. Especially for applications where sensor data are fused with coil signals for enhanced interpretation of images such as eye tracking [105], retrospective motion correction [37], correction of physiologically induced fields [104], [106], and magnetic field monitoring [55], [106], [107], synchronous acquisition of all data types is crucial.

Sensor and imaging data are used more and more for run-time adjustments of scanner operation. Examples include field stabilization [80], [108] and prospective motion correction based on active markers [43]–
[45], [96], and optical methods [40]–[42]. A particular type are navigator based methods which use either k-space [109], [110] or imaging data [111] rather than external tracking devices. All these methods profit from processing the underlying data with minimal latency.

Synchronous acquisition and real-time processing of an increasing diversity and quantity of data demand more flexibility, throughput, and an expansion of processing power, calling for a redesign of the traditional acquisition system.

To meet all these requirements, we propose the transition to reconfigurable and scalable hardware design, permitting an application tailored assembly of interfaces, real-time processing pipelines, and computational resources. We present an integrated implementation based on a single SoC, which offers sufficient throughput and hardware-based parallel processing power to meet real-time requirements for a large number of channels. The platform is equipped with versatile interfaces, which enable a synchronous acquisition of high throughput digital data streams. In form of fiber-optical modules, they address the trend towards heterogeneous and modular systems potentially with in-field operation.

We demonstrate the utility of the platform on the example of concurrent imaging and field sensing. The platform is used for fusion of coil data streams from two in-bore receivers presented in Chapter 2, their subsequent combination and compression to a smaller set of virtual coils according to [93]. The k-space trajectory is synchronously acquired using a field camera with in-bore digitization (Chapter 2) and computed in real-time on the platform. The trajectory is used for expanded encoding model based image reconstruction of single-shot readouts as commonly used in time-series fMRI [87]–[91]. Real-time coil compression and trajectory extraction demonstrate a high degree of data savings, which is a common bottleneck in time-series fMRI. Furthermore, multi-shot images were acquired to demonstrate high consistency and synchronicity of segmented data stemming from different sources.
3.2 Methods

3.2.1 Processing hardware platform

Figure 3.1: Hardware system architecture overview. The in-bore receiver system presented in Chapter 2 is part of a scalable, distributed processing architecture for multi-channel signal acquisition and real-time processing. In this chapter (Chapter 3), the re-configurable data collection and processing platform is focused.
An overview on the overall system architecture is given in Figure 3.1. We present a scalable and modular design concept to tackle application flexibility and data handling ability with increasing number of incoming channels. The system provides ample, configurable and distributed real-time computational power for advanced in-line data processing and low-latency applications.

Apart from flexible MR compatible in-bore or satellite units covered in Chapter 2, the out-field data collection and processing platform described in this chapter (Chapter 3) is an essential building block of the system architecture and is focused in the following.

The key component of the processing platform (Figure 3.2) is a SoC (Zynq-7045, Xilinx, San Jose, CA, USA). It combines a powerful processor (dual-core ARM Cortex-A9, Cambridge, UK) with a FPGA (Kintex, Xilinx, San Jose, CA, USA) in a single chip. In this configuration, it offers classical general-purpose computing power combined with a vast amount of parallel processing resources, which are necessary to process many-channel sensor and coil data. The FPGA consists of logic cells and digital macro blocks, which can be flexibly connected. The bit-stream defining these connections is loaded at power-up and can be freely updated with a new design, even on run-time. Just as any digital application specific integrated circuit (ASIC), the FPGA can fulfill arbitrary logic functionality, provided there are sufficient logic resources available. It is ideally suited to interface digital and mixed signal integrated circuits (ICs) of all kinds, as it supports a wide variety of pin standards, enables precise control of timing, and an easy implementation of protocols. Depending on the connected device, the interface standard and protocol can be flexibly changed, bringing software-like versatility to a hardware level. The FPGA is well suited for parallel execution of tasks on a large amount of data, rendering it a digital data hub and processing engine in a single chip. Furthermore, the processing is performed with minimal latency as is needed for MR applications with run-time adjustment of scanner operation.

The FPGA has multi-gigabit transceivers (GTX), which enable high-speed data transmission and high throughput. Those were connected to six fiber-optical modules (P1RX6B/P1TX6B, Inneos, Pleasanton, CA,
Figure 3.2: An overview of the platform hardware. Front (a) and back panel (c) and an inside view are shown (b). The core component of the platform is a SoC used to process a large number of channels from different sources concurrently and in real-time. To interface spectrometers and various sensors operating in-field, six high-throughput fiber-optical modules are used. The multi-mode fiber carries a low-jitter clock on one lane for synchronization of external modules with the platform. The clock is generated using a dual-PLL chip with external VCXO. A reference clock is provided from a high-quality OCXO, a GPS antenna or externally. A TTL-tolerant trigger interface is used to synchronize the platform operation with the scanner. The platform has a number of other useful interfaces such as SATA for data storage, Ethernet for remote access and data transmission, a high-throughput parallel interface, USB, SD card slot, and display port.
USA). As non-galvanic connections, they are very suitable for in-field modules. A single optical fiber combines four high-speed unidirectional (4 x 3.125 Gbps) and a bidirectional (1 x 1.2 Gbps) lane utilizing wavelength division multiplexing. A flexible usage of the available lanes is achieved utilizing programmable high-speed multiplexers (DS42MB200, Texas Instruments, Dallas, TX, USA) balancing between the number of available modules for data reception, throughput, and other interfaces.

Data processed in the SoC are either used to update the scanner operation or stored for image reconstruction. For this purpose, several options are available. Two fiber-optical modules mounted in transmit direction offer a low-latency connection and are therefore very suitable for real-time feedback as used in field stabilization [80]. Gigabit Ethernet allows standalone operation of the platform accessed via a console interface. A Serial Advanced Technology Attachment (SATA) interface serves the purpose of direct data storage on a solid-state drive. Alternatively, a parallel interface on the rear side of the platform offers full raw data throughput in the same amount as the four optical receive links. This interface can be used to connect to a console computer or allows multiple devices to be daisy-chained.

For synchronization of the connected in-field modules, each optical module provides a low-jitter clock signal on one lane. It is generated by an ultra-low phase noise clock generator based on a dual PLL (LMK04828, Texas Instruments, Dallas, TX, USA) with an external voltage-controlled crystal oscillator (VCXO) (CVHD-950, Crystek, Fort Myers, FL, USA). The reference clock is sourced from either a 10 MHz reference clock input, which is commonly used to synchronize the platform to the scanner clock, a global positioning system (GPS) disciplined oscillator (GPSDO) module with a GPS antenna, or an autonomous oven-controlled crystal oscillator (OCXO) (AOCJY6, Abracon, Spicewood, TX, USA). The option with OCXO provides a high-quality clock with 185 fs jitter integrated from 10 Hz to 1 MHz. It can be used to synchronize external devices through the reference clock output of the platform.

Furthermore, precision triggering and timing control is achieved by a delay matched trigger interface with four inputs and four outputs. Inputs
are transistor-transistor logic (TTL) tolerant and have programmable pull levels, and output & input impedance. On the hardware, an embedded ATxMEGA controller (Atmel) takes control of the high-speed routing multiplexers used for data-flow control, peripheral IO's and hardware supervision. All high-speed lanes of the optical modules are multiplexed to the 16 available GTX/GTH transceivers of the FPGA fabric or can optionally be re-directed such that two remote units can be directly connected, bypassing the main processing system. This enables the ability of two daughter modules to directly communicate e.g. for ultra-low latency feedback capabilities. The hardware platform includes two expansion connectors to accommodate optional hardware (e.g. clock feedback, see Chapter 4). An internal AC-DC converter connects to the systems DC-DC power modules and LDO power supplies to the mains power (90-230 V).

3.2.2 System-on-Chip

The SoC consists of two parts – the programmable system and programmable logic (FPGA) (Figure 3.3). The programmable system is a fully functional micro-controller system, which can be operated in a standalone fashion and run operating systems such as Linux. It consists of a processor with commonly used cores such as timer, interrupt controller, SPI, I2C, SDHC, UART, Ethernet, USB, and a SDRAM memory controller. They are connected with the processor via a system bus (AXI4-MM, ARM, Cambridge, UK). It is used to access the registers of the cores, which are seen as memory space. The functionality of the core can be controlled by a set of reads and writes, grouped in software functions and drivers for abstraction.

Programmable system and logic are connected with several such system buses (AXI4-MM) rendering the programmable logic a natural extension of the programmable system cores which can be flexibly designed. This approach enables high-level software control of the FPGA design and greater flexibility by moving parts of functionality to software. Additionally, it enables reuse of cores, use of third-party cores, and an easy and conformable integration, shortening the overall development time.
3.2. METHODS

Figure 3.3: The SoC consists of programmable system and logic parts. The programmable system is a dual core processor (ARM A9) with commonly used cores as SPI, I2C, USB, Ethernet, SDHC, timer, interrupt controller, Ethernet, and memory controller for SDRAM. The processor communicates with cores through a system bus (AXI4-MM), which is also used to access programmable logic (FPGA). It is thus a custom design extension of the programmable system cores. In this project it was used for data stream routing (blue) and signal processing (green) of a large number of data streams and channels concurrently and in real-time.

Within this project, cores have been designed to support high-speed optical interfaces of the platform, handling of the associated data streams, and large-scale real-time data processing, communication, and sequencing. A more detailed description of the designed cores is following.

High-Speed Data Link

To interface optical modules of the platform, a high-speed data link core has been developed. It is based on an Aurora core (Xilinx, San Jose, CA, USA) with an additional register interface for error monitoring, software reset and link initialization. The main purpose of the core is data serialization and de-serialization utilizing GTX transceivers, which operate at 3.125 GHz. The data are transferred to and from the core in a stream-like fashion using a parallel interface. It operates at a clock frequency of 156.25 MHz, which is used in the whole design as data processing clock. The interface is implemented in form of a simple point-to-point protocol (AXI4-Stream, ARM, Cambridge, UK). It has embedded data flow control, which is used in all signal-processing cores.
to raise error flags and interrupts to the processor if data streams are corrupted.

Memory Access

The bandwidth to the SDRAM memory amounts 51.2 Gbps, which in theory suffices for 10 stream interfaces at full bandwidth. However, it is a potential bottleneck for demanding applications with intensive data move to and from the SDRAM. For this reason, the main method for internal data exchange is realized using streaming interfaces. Instead, the SDRAM memory is used as a large buffer or as shared memory between the processor and programmable logic. It offers the possibility to combine the parallel processing power of the FPGA with complex operations in software, which are not suitable for a hardware implementation.

For this purpose, a memory access core has been developed. It is used to transfer data from the streaming architecture to memory and the other way around. It has four master and slave stream interfaces (AXI4-Stream) and a master interface to a memory mapped bus (AXI4-MM master). The latter is connected to the programmable system counterpart (AXI4-MM), which has access to the memory controller interfacing SDRAM. Memory locations, data selection and block sizes can be defined by register settings and an interrupt can be raised upon completion of a block transfer for flow control.

Stream Interconnect

To enable concatenation of signal processing operations and flexible routing of data streams, a stream interconnect core has been developed. It connects up to 32 slave and master stream (AXI4-Stream) interfaces and it is a central point where all data passes through. Consequently, it is a critical part, which has high impact on placement and routing of the whole FPGA design. It utilizes several pipelining stages and a clever multiplexer architecture for easier timing closure of the digital design.
3.2. METHODS

Stream Fusion

For joint processing and storage of common source data, a stream fusion core has been developed. It fuses up to four data streams into a single one with prior channel selection. This covers the case of all four optical modules connected to the same data source, e.g. four in-bore receivers connected to a 64-channel coil.

Parallel Interface

The parallel interface core offers high throughput sufficient to transfer data from all optical modules to the host computer at full bandwidth. As GTX resources are scarce and rather used to interface the optical modules of the platform, general-purpose pins have been used. They can operate at a frequency of up to 1 GHz, which helped to minimize the lane count. Data and control signals of the four stream interfaces (AXI4-Stream) are 5b6b encoded for DC balancing and serialized into a parallel data stream operating at 937.5 MHz. It is transferred via two cable connectors placed on the backside of the platform.

Phase and Magnitude Extraction

Field sensors based on NMR as NMR active markers [43], [44], [95], [96] and field monitoring probes [35], [45], [55], [56], [97], [98] rely on measurements of free-induction-decay (FID) signals generated in pulsed NMR experiments. The phase of such a FID is directly proportional to the temporal integral of the magnetic field at the sensor location, which is the desired value for k-space encoding. The first processing step in field sensing is therefore evaluation of the signal phase for which a dedicated core has been developed.

It is based on a CORDIC in circular vectoring mode, which by applying a set of elementary rotations, shift and add operations, successively approximates the phase and magnitude of a complex signal [72]–[74]. For maximal numerical precision and full throughput, it performs 32 iterations in pipelined fashion. Magnitude and phase results are split
into two separate data streams (AXI4-Stream) since further processing usually differs for the two data types.

The core has a number of optional additional features as follows. To directly obtain the magnetic field values instead of the temporal integral, the phase can be differentiated. For further processing, it can be unwrapped. To accommodate for phase wraps and to avoid contamination with quantization noise, the representation of unwrapped phase is extended to 64 bits and output as upper and lower 32-bit words at the stream interface (AXI4-Stream). In this way, a unified 32-bit interface can be used throughout the design. Arithmetic operations require minor modifications to support both bit widths [112], [113], which are outlined in more detail in the FIR Filtering and Decimation section below. Furthermore, for off-resonance correction a linear phase can be added to the unwrapped phase, which is calculated by accumulating a phase increment for each input channel.

**FIR Filtering and Decimation**

Since the phase evolution of field probe signals have a considerably lower bandwidth than their complex raw data counterpart, which typically even exceeds imaging signals due to their positioning at the edge of the imaging volume, a decimation filter can be applied to reduce the amount of data without loss of information.

To cover a wide spectrum of applications and to keep the resource utilization at minimum, a customizable FIR filter architecture was implemented. The number of MAC units and input buffers can therefore be selected prior to synthesis. The internal structure of the core shares the available resources balancing the load at each input buffer – MAC pair equally. Each new incoming sample is stored in the subsequent input buffer with the restriction that samples corresponding to the same channel are stored in the same buffer (Figure 3.4). During convolution, the input buffer read address \((Rd_{Addr})\) is decremented from the address of the newest sample by a factor \(Rd_{AddrDec}\) equal to the ratio of the
Figure 3.4: Internal architecture of (a) FIR decimator and (b) matrix-vector multiplier core. (a) Incoming samples to the FIR decimator are stored in a buffer, such that the load is equally balanced on all input buffer-MAC pairs. Convolution is performed as a series of multiply and add operations in the MAC unit. Samples belonging to the same channel are buffered in the same input buffer. Numbers under the input buffer array and over the coefficient buffer indicate the order of execution. Please note that the input buffers are addressed with an address decrement of two as there are 7 channels and 4 input buffer-MAC pairs in the example. Decimation and an arbitrary number of channels can be processed with the proposed architecture without limitations. Only addressing of the input buffer has to be adjusted and is programmed via software. (b) Matrix-vector multiplication is executed as a series of row-vector multiplications. Depending on the number of MAC units, they are executed in parallel until a whole matrix-vector multiplication has been performed. Matrix size is programmable via software whereas the maximum size is given with the size of the coefficient buffer array.
number of input channels ($N_{\text{Channels}}$) to the number of MACs ($N_{\text{MACs}}$) rounded down to the first integer:

$$Rd_{\text{AddrDec}} = \left\lfloor \frac{N_{\text{Channels}}}{N_{\text{MACs}}} \right\rfloor$$ (3.1)

If there are more channels to process than MACs ($N_{\text{Channels}} > N_{\text{MACs}}$), the start read address ($Rd_{\text{AddrStart}}$) is decremented by one and the following convolution is performed for the subsequent channel. After all channels have been processed, the next convolution is performed after $D$ incoming samples, where $D$ corresponds to the decimation factor set via register. All input buffers and MACs operate in parallel and share the same coefficients buffer. Please note that $N_{\text{Channels}}$ is not implementation limited and it is sufficient to set it via register for accurate $Rd_{\text{Addr}}$ calculation. For a given implementation, the maximal number of FIR coefficients ($N_{\text{Coeffs}}$) equals to:

$$N_{\text{Coeffs}} = \left\lfloor \frac{f_{\text{sys}}}{f_{\text{in}}} \cdot D \cdot \left( \left\lceil \frac{N_{\text{Channels}}}{N_{\text{MACs}}} \right\rceil \right)^{-1} \right\rfloor$$ (3.2)

where $f_{\text{in}}$ is the sampling rate of the incoming samples and $f_{\text{sys}}$ the clock frequency the FIR decimator is operating on. The lower bound of $N_{\text{MACs}}$ can then be determined according to the worst-case parameters for $N_{\text{Channels}}$ of the pre-decimated data ($f_{\text{in}}/f_{\text{sys}}$), the target $D$ and the desired filter length ($N_{\text{Coeffs}}$):

$$N_{\text{MACs}} = \left\lceil \frac{f_{\text{in}}}{f_{\text{sys}}} \cdot D^{-1} \cdot N_{\text{Coeffs}} \cdot N_{\text{Channels}} \right\rceil$$ (3.3)

To avoid overflow, the input buffer needs a minimum size ($N_{\text{BufferSize}}$) of:

$$N_{\text{BufferSize}} = N_{\text{Coeffs}} \cdot \left\lceil \frac{N_{\text{Channels}}}{N_{\text{MACs}}} \right\rceil$$ (3.4)

The core supports single and double precision fixed-point data, which are treated as separate channels with slight modifications as follows. For proper multiplication of double precision data, the upper word has a one-bit sign extension, whereas the lower word is extended with zero prior standard signed multiplication with the filter coefficients. To support both precision types with the same implementation, the single precision
data are always sign extended. After convolution, the MAC accumulator values corresponding to a double precision word are added with a 32-bit shift for the upper word and rounded to 64 bit. The result is output as an upper and lower 32-bit word on the stream interface (AXI4-Stream). A single register bit is sufficient to determine the precision. Double precision arithmetic is known from the first days of digital computer and first mentioned by von Neumann in 1947 [112], [113]. An illustration of double precision arithmetic along with single precision support is given in Figure 3.5.

**Matrix-vector multiplication**

Coil arrays and many sensor systems deliver data in form of multiple channels. For further processing, they often require a linear transformation step in the channel dimension, which is a matrix-vector multiplication. For this purpose, a dedicated core has been developed. In this work, it is applied to field monitoring where probe phases are fitted onto a set of spherical harmonics in order to obtain k-space trajectories and higher order k-space terms [55]. Further, it is used for coil combination and compression in time-domain [93], [94].

The matrix-vector multiplier uses a similar internal structure as the FIR decimator with shared MACs operating in parallel. However, it has only one input-buffer for the input vectors and several coefficient buffers for each row of the matrix (Figure 3.4). Each MAC performs a multiplication of the input vector with a row of the matrix. If there are more rows than MACs ($N_{Rows} > N_{MACs}$), the multiplication is first performed with the next set of rows until the matrix-vector multiplication has been completed. The size of the matrix is set in run-time via register, whereas the maximal size of the matrix is determined via generic. The lower bound for the necessary $N_{MACs}$ is given by:

$$N_{MACs} = \left\lceil \frac{f_{in}}{f_{sys}} \cdot N_{Columns} \cdot N_{Rows} \right\rceil$$

(3.5)

The core supports double precision data utilizing the same principles outlined in the FIR decimator section.
Figure 3.5: Fixed point arithmetic for single and double precision data. The bit-widths are indicated with a common notation and rounding is omitted for simplicity. Sign bits are indicated with S, whereas ext signifies extension to prevent overflow. MSBs, LSBs and zeros are used in a similar fashion to differentiate the actual word from bits resulting from an arithmetic operation. Examples are given for 32-bit (a) single and (b) double precision data and 25-bit coefficients as used in the signal processing cores.
Sequencing

Operations like unwrapping, off-resonance correction, and decimation require a reset for each new interleave. This is as well the case for spectrometer filters and sensors operating in-field.

To accommodate four independent data streams from four optical modules, a trigger generator with four internal pulse generators has been designed. The pulse generators can be flexibly triggered by a logical combination of external, software and trigger generator outputs. However, they are mainly used to synchronize the data acquisition and processing with the scanner operation. The pulses are source for 32 trigger outputs connected to external ports and cores, which need synchronization. Optionally, interrupts are raised on pulse triggers to synchronize software routines called at the beginning of each new interleave.

3.2.3 Software control

All implemented cores and the data processing pipeline can be controlled in real-time by a control software with appropriate user interface. By defining the connected hardware blocks, the software automatically configures the whole hardware tree, from receiver gain module amplification to matrix vector multiplier coefficients.

3.2.4 Experimental setup

All experiments were performed on a 7 T human whole-body MRI system (Philips Achieva, Philips Healthcare, Best, The Netherlands). Data were collected from a healthy volunteer according to the applicable ethics approval using a concurrent imaging and magnetic field monitoring setup as in [91], [114]. It consists of a commercially available quadrature-transmit coil surrounding a 32-channel head receive array (Nova Medical, Wilmington, MA, USA) used for imaging. In between, an array of 16 NMR field probes [35], [56], [97] is mounted on a laser-sintered polyamide frame and integrated in the head setup. The probes have a hexafluorobenzene filling, an inner capillary diameter of 0.8 mm, $T_1 =$
86 ms, and $T_2^* = 24$ ms. They were connected to a standalone version of the T/R front-end with integrated pre-amplifiers as presented in [56] and placed on top of the coil pre-amplifier boxes in the scanner bore. For excitation with short frequency-modulated pulses, a standalone version of the DDS module presented in [56] was used. The field monitoring front-end was connected to an in-field receiver module presented in chapter 2 which was used to acquire the probe signals and to control the DDS and T/R switch timing. Two additional 16-channel in-field receivers were used to acquire the 32 coil signals, which were passively split between the scanner spectrometer and the in-field receivers at the panel under the scanner bore. This allowed for convenient scan planning and full scanner preparations without notable impact on the overall noise figure of the receive chains since the splitting was done after the coil pre-amplifiers.

The in-field receiver modules were fiber-optically attached to the platform occupying three of four available receive modules. Through the back-channel of the optical link they were frequency locked to the processing platform clock providing full synchronicity to the measurements without the need for data re-sampling or delay calibration.

Data acquisitions were started on an external trigger signal from the MRI scanner provided to the processing platform sequencing core and forwarded to the in-field receivers. The trigger signals were sampled using the main system clock of 156.25 MHz at the in-field modules and resulting in a synchronization resolution of 6.4 ns.

The in-field receivers delivered raw data streams, which were concurrently processed on the platform SoC. The coil data streams were fused and compressed to a smaller set of virtual coils, whereas the probe data stream was used to calculate the k-space trajectory. Along the processed data, raw coil and probe data streams were output on the parallel interface for comparison. An overview of the stream-based signal processing architecture is given in Figure 3.6. A detailed description of the applied signal processing steps follows.

For joint processing and storage of all coil signals, two imaging streams were fused to a 32-channel stream. They were combined and compressed according to [93] using a matrix-vector multiplier which was
Figure 3.6: Data streaming and signal processing architecture. The coil signals are received through two high-speed data link cores, routed to a stream fusion, and subsequently to a matrix-vector multiplier core for coil compression (all cores are shown in blue). Along coil compression data, raw coil signals are routed to the parallel interface for comparison. The probe signals were first routed to a matrix-vector multiplier for decoupling followed by a core for phase extraction, unwrapping, and off-resonance correction. A second matrix-vector multiplier is used for phase to trajectory transformation, and finally a FIR decimator core for trajectory smoothing. Smoothed trajectory and raw probe signals are routed to the parallel interface for storage and comparison. All cores used for trajectory processing are shown in green.
configured to output 12 channels. The compression matrix was reloaded for each new slice for improved noise performance. The reloading was implemented as software routine, which is configurable in a similar way to a core and triggered by an interrupt from the sequencing core.

The probe signals were first fed to a matrix-vector multiplier core for decoupling according to [106] and described in more detail in [115]. It was followed by phase extraction, unwrapping and off-resonance correction implemented in the phase extraction core. Compared to the raw probe signals, the gradient fields do not exceed 50 kHz in bandwidth [56], [116]. Hence, the unwrapped phase was passed through a FIR filter designed to allow 16-fold decimation and consequently 12 dB SNR improvement. The final step for trajectory evaluation was another matrix-vector multiplication used to transform the probe phases to a second-order spherical-harmonic field model. The model fit was based on probe positions which were evaluated in a calibration scan as described in [55], [114], [91].

3.2.5 Imaging sequences

Two scanning protocols were used, segmented multi-shot spirals for anatomical imaging as in [114] and single-shot spirals as used in time-series fMRI [87]–[91]. Both spiral trajectories were Archimedean with center-out direction, incorporated in a multi-slice 2D gradient echo sequence with a fat suppression SPIR module preceding slice selection [117]. In total, 9 transverse slices have been acquired with a volume TR of 1s. The slice thickness and inter-slice gap were 2.00 mm and 1.51 mm respectively. The readouts had a nominal TE of 18 ms for T$_2^*$ weighting and covered an isotropic FOV of 230 mm. The spiral gradient waveforms utilized full gradient strength and slew-rate of 31 mT m$^{-1}$ and 200 mT m$^{-1}$ ms$^{-1}$ respectively for minimal readout duration [91], [114], [118]. For the anatomical images, 24 interleaves with a duration of 29 ms were acquired for full k-space coverage and a nominal in-plane resolution of 0.4 mm. The single-shot spirals were under-sampled by a factor of 4 resulting in 0.8 mm in-plane resolution for 56 ms long readouts.
3.2. METHODS

For calculation of static off-resonance and coil sensitivity maps, a fully sampled Cartesian gradient-echo scan with same geometry as the spiral scans and a 1 mm in-plane resolution was acquired.

All scans were concurrently monitored. The field probe excitation was performed shortly before the start of the readout gradients to allow for longest possible probe lifetime. Every 3rd slice was monitored resulting in a spacing of >300 ms for T1 recovery of the probes. It is still sufficient to account for breathing-induced field changes as described in [114]. The acquisitions of imaging data were started earlier to include the RF pulse, which was used for phase synchronization of the coil channels.

3.2.6 Image reconstruction

Image reconstruction was based on an expanded encoding model including static off-resonance and coil sensitivity maps along with zero and first order k-space trajectories [55], [91], [114].

Higher-order k-space trajectories were neglected in the reconstruction step due to reconstruction speed and negligible effect [91], [114]. Concomitant field effects have been modeled from the monitored first-order terms and retrospective correction was applied [119].

The Cartesian gradient-echo scan was reconstructed in the same way but without static off-resonance and sensitivity maps. Coil sensitivity maps were obtained from the first-echo images dividing each single-coil image by the root-sum-of-squares image of the array. Subsequently, coil compression matrices were calculated according to [93] in a slice-wise fashion. Static off-resonance maps were calculated by pixel-wise fitting the phase evolution across the echoes. Both maps were interpolated and smoothed as described in [91].

Image reconstruction was performed by inversion of the encoding model using an extension of the iterative conjugate-gradient SENSE algorithm [91], [114], [120].
3.3 Results

Figure 3.7a shows the trajectory evolution for zero, first and second order spherical harmonic components for a single interleave of a monitored multi-shot spiral scan. Zero and first order components were used along with static off-resonance maps in image reconstruction, while the second order components were neglected due to their small magnitude.

The corresponding $T_2^*$-weighted multi-shot spiral images with 0.4 mm in-plane resolution are shown in Figure 3.8. The images exhibit competitive quality, which is high in all slices including the lower ones. A high degree of anatomical detail with very fine structures, particularly small vessels can be observed. A hypo-intense feature is visible in the upper slices between the frontal lobes. It has been identified as a calcification of the falx cerebri, which is a non-pathogenic variation within the healthy population. The absence of typical artifacts common for spiral imaging, as blurring and distortion, affirms high consistency among the spiral interleaves. This confirms that trajectory and both coil channel groups have been acquired synchronously without conspicuous jitter among the acquisition starts. Furthermore, spiral acquisitions are known to be highly sensitive for static off-resonances [91], [114]. They were calculated from a separate scan and used throughout the whole scan session, which took approximately one hour, confirming high overall stability.
3.3. RESULTS

Figure 3.7: (a) Trajectory of a single interleave of a fully sampled multi-shot spiral with outward direction and a TE of 18 ms. Maximum phase excursion in a sphere of 10 cm radius (“max rad”) is shown. Spherical harmonic orders are separated and shown in different graphs (rows). The trajectory was calculated on the platform FPGA on basis of 16 field monitoring probes fitted to a second order spherical harmonic model. Concomitant fields effects were modeled from the monitored first-order terms and retrospectively subtracted. (b) Trajectory of a four-fold under-sampled single-shot spiral with outward direction and a TE of 18 ms calculated in the same way as in (a). (c) The difference between single-shot spiral trajectories calculated in real-time and retrospectively. It can be observed that the error has similarity with the dominant 1st order k-space terms. This is due the finite precision of the matrix-vector multiplication in the phase to k-space transformation step. This is also the reason for the slight differences in the trends of the error along with finite precision of off-resonance correction of the probes.
Figure 3.8: T₂-weighted (TE = 18 ms) outward spiral acquisition with 0.4 mm in-plane resolution and 2 mm slice thickness. 9 Slices (a-i) have been acquired with 24 interleaves covering the k-space with full sampling. The image encoding model incorporated static off-resonance maps along with δ\textsuperscript{th} and 1\textsuperscript{st} k-space trajectory terms evaluated in the platform FPGA in real-time. The images were reconstructed on using eight virtual coils calculated on basis of SENSE maps according to [93].
Figure 3.9 shows single-shot images with 0.8 mm in-plane resolution and 2 mm slice thickness and Figure 3.7.b the corresponding trajectory. The images exhibit \( T_2^* \) contrast as typically utilized in fMRI time-series acquisitions [87]–[91]. Distinction between gray, white matter and cerebrospinal fluid (CSF) is very clear and cortex boundaries are to a large extent sharp despite very long readouts. Especially at high field strength, long readouts reveal intra-voxel dephasing which is particularly prominent in the frontal lobes and close to the skull, where the susceptibility gradients are at their largest. The images appear without significant blurring and distortion, which were traditionally associated with spiral readouts.

Figure 3.7c shows the error of the real-time calculated trajectory. It is compared to the retrospectively calculated trajectory on basis of raw probe data. The relative difference is in the order of \( 10^{-6} \) compared to the first order components and \( 10^{-4} \) compared to 0th and 2nd order spherical harmonics. The error has high similarity to the 1st order k-terms identifying the finite precision of the matrix-vector multiplier coefficients in the phase to k-space transformation step as the main source of error. Furthermore, small differences in the trends of the different components could indicate a residual off-resonance error due to 32-bit precision along with the matrix-vector multiplier coefficients. It can be observed that the error of some components is larger compared to the others, mainly \( x, y \) and \( z^2 \). This is due to the fact that concomitant field effects were estimated from the first-order components [119] and retrospectively corrected scaling the total error accordingly.

Figure 3.10 shows a slice of a single-shot spiral along with the error images due to coil compression and real-time coil compression with the according scaling. Coil compression comes at the cost of 6.15\% SNR for 8 virtual coils and 3.07\% for 12 coils. The difference between retrospective and real-time coil compression was quantified to be below \( 10^{-6} \) for a reconstruction performed with 20 iteration for both images.
Figure 3.9: Four-fold under-sampled single-shot spiral with outward direction and a TE of 18 ms. 9 slices (a-i) have been acquired with 0.8 mm in-plane resolution and 2 mm slice thickness. The image encoding model incorporated static off-resonance maps along with $d^{th}$ and $1^{st}$ k-space trajectory terms evaluated in the platform FPGA in real-time. The images were reconstructed using eight virtual coils calculated on basis of SENSE maps according to [93].
3.4 Discussion

An MRI platform with increased data handling and real-time processing capabilities was presented. It employs the concept of re-configurable hardware, permitting a tailored assembly of interfaces and real-time processing pipelines which can be flexibly assigned via software. These real-time computing capabilities are essential for enabling very large channel count acquisitions and to keep data flow, storage and reconstruction tractable. The platform has versatile interfaces in form of fiber-optical modules with embedded synchronization, suitable for spectrometers and sensors operating in-field.

It was applied for concurrent imaging and field monitoring with real-time coil compression and trajectory extraction. It has been shown that data fusion on a common time base enables a consistent combination of channels and data types. Compression of imaging data and real-time processing of field probe data are practically equivalent to conventional offline computation. The observed discrepancies are due to fixed-point
arithmetic. The numerical precision was designed to render the presented cases satisfying. Compared to floating-point it offers particularly economical hardware implementation and is commonly used in hardware signal processing.

The applied compression and sensor processing enabled substantial data reduction. Coil compression was implemented to use 12 virtual coils out of 32 physical array elements with a 3.07% SNR penalty. For 8 virtual coils corresponding to a 4-fold data reduction, the SNR penalty amounted to 6.15%. It is important to note that the noise performance can be considerably improved when coil compression is based on regions of interests (ROIs) targeting specific anatomies [93], [94]. It is becoming an ever more important technique as number of array elements increase [20], [21], [82].

Trajectory extraction was based on standard processing and comes without signal quality deterioration. It enables significant data reduction as gradient fields typically have a much lower bandwidths compared to raw sensor data [55], [56]. The bandwidth depends on the application and is in the order of tens of kHz. In this work, 16-fold decimation was employed for a bandwidth of 50 kHz. If only zero and first order k-space terms are of interest, as with gridding-based image reconstruction employed in this work, further data reduction is possible. With 16 field probes, an overall 64-fold data reduction was achieved. Compression and immediate processing of coil and sensor data are particularly important for data intensive tasks as time-series fMRI [87]–[91].

The platform has a high-quality clock and provides various options for synchronization of connected modules. It is equipped with versatile optical interfaces and readily operable with in-bore receivers and sensors [83], [85], [86], [121]. In this work, it was used for a combination of MR imaging and field sensor data. However, it is ready for any digital sensor signals ranging from physiology as cardiac and breathing sensors [100]–[104] to more complex sensors as cameras for motion correction [37], [40]–[42] or eye tracking [105]. Moreover, FPGAs as integrated here are commonly used for video processing including computer vision tasks such as eye tracking [105]. For this purpose, modular additions to
the signal processing and stream handling architecture of the SoC are straightforward to integrate. The architecture covers frequently used operations and can be extended as well with asynchronous computing on a processor for complex algorithms or fast prototyping.

Promising future applications cover low-latency field stabilization, dynamic slice-wise shimming and motion correction for which the platform is readily applicable. Low-latency calculation could significantly contribute to mitigate physiologically and motion induced signals beyond breathing and cardiac activity. Real-time and parallel processing prospects an unrestrained integration of all feedback modalities along with retrospective correction for effects beyond the feedback loop bandwidth. The best performance in terms of precision and latency is expected to be achieved if console and sequence generation functionality move to the platform for which it is readily applicable.
Chapter 4

In-Bore Clocking and Gradient Influence Correction

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4.1 Introduction

In a recent trend to further increase the number of receive channels in MRI, high-dynamic range receive electronics are placed in-bore [121] or even on-coil [122]. These receivers digitize and encode the signals for efficient digital transmission on a non-galvanic medium, removing the need for ponderous and coupling cables used traditionally. Cables have been replaced by wireless [57], [59], [60], [123] or optical links [32], [63], [121], overcoming RF related difficulties, re-establishing range of motion and enabling more manageable coil engineering.

However, signal digitization in-bore requires sensitive electronics to be operated in a harsh EM environment, which has several pitfalls that may limit achievable performance [124] & Chapter 2.

One of the factors is the quality of the oscillator signal needed by the analog frontend. All common receiver topologies require a precise clock at the frequency-conversion stage either in form of a local oscillator (LO) for analog mixing or as encoding clock for the ADC.

Considering that a strong static magnetic field with fast switching gradients and powerful RF pulses pervade the electronics, careful clock system design is required for in-bore receivers. Scanner EMI can hamper local oscillators or clock signals fed from an external source. The consequent signal alterations give rise to image distortions or measurement errors when acquiring sensor data, such as field probes [35], [125].

In this work, we explore clocking in-bore, revisit clock generation and compare performance under gradients of microelectromechanical systems (MEMS) and crystal oscillator devices. It is confirmed that gradient switching modulates state-of-the-art clock sources and demonstrated that gradient action hampers VCO based PLL circuitry, potentially reducing SNR or causing image distortions.

To tackle these detriments, in case of an optically connected 16-channel in-bore receiver (see Chapter 2 & ref [121]), we propose a method for correcting scanner EMI induced signal modulations and clock source imperfections.
Based on a feedback mechanism, successful compensation of gradient influence on the sampling clock of the in-bore receiver system is demonstrated. For that, a dedicated hardware was developed with real-time implementation of the correction scheme in an FPGA.

With the suggested mechanism in place, performance of synchronous sampling in-bore receivers is on par with traditional out-field systems and is suitable even for high dynamic range in-field acquisition of field monitoring data.

The presented compensation mechanism is especially valuable for high bandwidth in-bore receivers but can be of interest in applications where synchronous acquisition of multiple, distributed receive channels is hindered by a harsh EM environment.
4.2 Theory/Background

In the following, a short overview on the importance of the oscillator signal in a receiver in general is given and potential vulnerabilities of oscillator in-bore operation are highlighted.

4.2.1 Clock noise

A real oscillator signal \( s_{\text{clk}} \), contains a certain amount of phase \( \phi(t) \) and amplitude \( \varepsilon(t) \) error. This signal has a nominal radian frequency \( \omega_{\text{clk}} \) and an amplitude \( A = A_0 + \varepsilon(t) \). With the relation \( \text{d}\phi = \text{d}\omega \cdot \text{d}t \), and the timing error referred to as jitter \( \tilde{t} \), this is expressed as

\[
s_{\text{clk}}(t) = A \sin(\omega_{\text{clk}} t + \phi(t)) \\
= A \sin(\omega_{\text{clk}}(t + \tilde{t}(t))) \\
= A \sin(\omega_{\text{clk}} t) \cdot \cos(\omega_{\text{clk}} \tilde{t}(t)) + A \cos(\omega_{\text{clk}} t) \cdot \sin(\omega_{\text{clk}} \tilde{t}(t)).
\]  

(4.1)

Phase-noise is generally defined as the frequency representation of this random fluctuation of the phase of a waveform [126]. A commonly used definition relates the actual (sideband) power of phase fluctuations and the nominal or average frequency \( (f_0) \) and is called \( L(f) \), in units of dBc/Hz. \( L(f) \) is expressed as a normalized spectral density where

\[
\int_{-\infty}^{\infty} L(f) \, df = 1
\]

and

\[
L(f) = \frac{\text{phase noise power in 1 Hz bandwidth at offset } f}{\text{carrier power } (f_0)}.
\]

Plotting \( L(f) \) shows a combination of discrete-frequency noise components with their own spectral densities (e.g. power-line or vibration frequencies) and power-law noise processes which can generally be classified into five categories [127] as shown in Figure 4.1.
### 4.2. THEORY/BACKGROUND

#### Noise processes

<table>
<thead>
<tr>
<th>Noise Process</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>White phase</strong></td>
<td>Thermal noise (Johnson noise)</td>
</tr>
<tr>
<td></td>
<td>kT buffer amplifier noise, resistor noise and Shott noise</td>
</tr>
<tr>
<td><strong>Flicker phase</strong></td>
<td>Pink noise</td>
</tr>
<tr>
<td></td>
<td>buffer amplifier flicker noise</td>
</tr>
<tr>
<td><strong>White frequency</strong></td>
<td>Carrier noise</td>
</tr>
<tr>
<td></td>
<td>Crystal RLC noise</td>
</tr>
<tr>
<td><strong>Flicker frequency</strong></td>
<td>Intermodulation of White Frequency (Carrier noise)</td>
</tr>
<tr>
<td></td>
<td>and Flicker phase (transistor noise), particularly within the oscillator loop</td>
</tr>
<tr>
<td><strong>Random walk</strong></td>
<td>Intrinsic noise sources within Quartz and electrode structures</td>
</tr>
<tr>
<td></td>
<td>External effects by environmental changes e.g. mechanical shock, temperature changes, vibration, etc.</td>
</tr>
</tbody>
</table>

*Figure 4.1: Noise spectral density and the five independent noise processes [127]*
Jitter is the equivalent of phase noise in the time domain and can be calculated as the integral of spectral phase density $\mathcal{L}(f)$ with respect to frequency between two limits, typically the signal bandwidth of interest. The result is frequency independent and written as

$$\sigma = \frac{1}{2\pi f_0} \sqrt{\int_{\frac{T_{\text{meas}}}{2}}^{f_{\text{BW}}/2} 2\mathcal{L}(f) \, df},$$

where $\sigma$ is RMS jitter in seconds. To convert to RMS phase jitter in radians, $\sigma$ is multiplied by the fundamental frequency: $\sigma \cdot 2\pi f_0$.

For oscillators phase noise is usually specified from a kHz range (12 kHz is a typical value) to large offset frequencies because for communication systems the signals reside further from the carrier and "for most modulations including 3G, losing signal energy around DC is acceptable" [128]. In contrast, MRI has different requirements because a stable acquisition phase is part of the image encoding principle. Therefore, the total observation time ($T_{\text{meas}}$) governs the lower bound of integration. This can have a significant impact on performance as for most oscillators close-in phase noise dominates.

### 4.2.2 Oscillator signal in a receiver

To recall the importance of the clock in a receiver system let us examine two essential operations: frequency translation by multiplication and sampling. Any receiver system performs either or both, often even multiple times.
Clock errors in a mixer

In a simplistic model of an ideal frequency translating mixing stage, an input signal \( s(t) \) is multiplied with a clock or LO signal \( s_{\text{clk}}(t) \) (eq. 4.1):

\[
u(t) = s(t) \cdot s_{\text{clk}}(t) = s(t) \cdot \{A \sin(\omega_{\text{clk}} t) \cdot \cos(\omega_{\text{clk}} \tilde{t}(t)) + A \cos(\omega_{\text{clk}} t) \cdot \sin(\omega_{\text{clk}} \tilde{t}(t))\}
\]

\[
\approx s(t) \cdot A \sin(\omega_{\text{clk}} t) + s(t) \cdot A \cos(\omega_{\text{clk}} t) \cdot \tilde{t}(t)
\]

\[
\approx \underbrace{s(t) \cdot A \sin(\omega_{\text{clk}} t)}_{\text{wanted}} + \underbrace{s(t) \cdot A \cos(\omega_{\text{clk}} t) \cdot \tilde{\phi}(t)}_{\text{unwanted}}
\]

(4.2)

The first term is the desired mixing product, the second term the unwanted product due to LO phase noise [129]. On the mixer output, in the frequency domain, the phase spectrum of the clock signal appears convoluted with the frequency-translated input. In an otherwise ideal receiver, this phase error of the clock limits the receiver sensitivity.

Clock errors in ADC sampling

Similarly, the ideal sampler in an ADC is a binary switch where the sampling time is determined solely by the instant when the clock goes high (or low) and passes a fixed threshold. A signal \( s(t) \) sampled at the rate \( \frac{1}{T} \) is \( s[n] = s(nT) \), where square brackets represent discrete values.

Real sampling of this signal with an encoding clock with timing error \( \tilde{t} \), produces sample misplacements that can be expressed as:

\[
\tilde{s}[n] = s(nT + \tilde{t}(nT))
\]

Without loss of generality, assume a sinusoidal input signal:

\[
s(t) = C \sin(\omega_{\text{in}} t).
\]
By sampling this becomes:

\[
    u[n] = C \sin(\omega_{in} (nT + \tilde{t}(nT))) \\
    = C \sin(\omega_{in} nT) \cdot \cos(\omega_{in} \tilde{t}(nT)) + C \cos(\omega_{in} nT) \cdot \sin(\omega_{in} \tilde{t}(nT)) \\
    \approx C \sin(\omega_{in} nT) \underbrace{+ \cos(\omega_{in} nT) \cdot \omega_{in} \cdot \tilde{t}(nT)}_{\text{wanted}} + \underbrace{C \cos(\omega_{in} nT)}_{\text{unwanted}} \cdot \omega_{in} \cdot \tilde{t}(nT)
\]

(4.3)

Again, the first term is the desired sampled signal, the second term the unwanted jitter induced product.

Notice the three differences to analog mixing: 1) small errors in sampling clock amplitude have no effect on sample misplacement. This is because of the sensible approximation of an ideal sampling switch [130]. The amplitude has to be large enough to not degrade the ADC aperture jitter. 2) the unwanted product scales with input frequency \( \omega_{in} \) as opposed to the LO frequency \( \omega_{clk} \) in a mixer. This is substantial and arguably the biggest drawback for high input frequency direct sampling architectures. 3) the sampling frequency has no direct influence on the introduced error.

The spectrum of the phase noise \( \omega_{in} \cdot \tilde{t}(nT) = \tilde{\phi} \leftrightarrow \Phi(f) \) shows up in the frequency domain as a convolution with the noise-free input signal \( s(t) \) [130]. Phase noise appears as sidebands around the center frequency \( \omega_{in} \).

Clearly, timing infidelity influences both sampling and mixing (Figure 4.2).

4.2.3 SNR limit in direct sampling receivers

In this work a receiver based on a direct sub-sampling architecture [69] was used. Today's broadband ADCs enable direct sampling of input signals with frequencies of several hundred megahertz (MHz). Direct undersampling with intentional aliasing is used to leverage this broad input bandwidth with a limited sampling frequency [131]. Besides having many advantages, this technique is especially prone to clock jitter errors, as shown above. In addition to being proportional to the input frequency, noise is aliased for each of the Nyquist zones recorded [132].
The dynamic range and SNR of such a receiver is governed by the phase noise of the encoding clock with a theoretical limit of \([77]\):

$$\text{SNR}_{\text{jitter}} = -20\log_{10}(\omega_{\text{in}} \cdot \sigma)$$

in dB fullscale (dBFS), where \(\omega_{\text{in}}\) is the input frequency and \(\sigma\) the jitter in RMS seconds. An in-depth analysis of ADC clocking jitter is given in \([77]\), \([129]\), \([133]\).

### 4.2.4 In-bore operation of oscillators

As alternative to the undesired cables, clock transmission using optical fibers may add excessive jitter. Similarly, wireless clock transmission suffers from multipath distortions. Therefore, direct clocking is not an option for successful in-bore signal acquisition. It relies on a local field-exposed oscillator, either stand-alone or as part of a PLL circuit. PLLs are useful to improve the phase noise of an externally sourced reference clock and are essential for long-term stable and synchronous acquisition in case of multiple receivers.
Fundamentally, oscillators rely on a resonating element for frequency determination and an active element to overcome dissipative losses and sustain oscillation. Both elements give rise to different types of noise [134]. The resonating element is the primary noise source close to the carrier, while the oscillator sustaining circuitry is the primary source of noise far from the carrier (Figure 4.1). Resonating elements oscillate at fixed frequencies and include, but are not limited to, electronic delay lines, LC tanks, dielectric resonators, electromechanical elements such as crystals and MEMS, and range to the most precise ytterbium atomic timing standards relying on ion energy transitions [135]. Combining these central elements with additional compensatory circuitry at discretionary complexity helps improving stability and environmental independence.

For in-bore use, however, a compact and non-magnetic system is desirable, precluding the use of large supplementary circuitry such as an active oven for temperature stabilization or a suspension for vibration compensation. Broadband application requires the oscillation frequency of a receiver to be tunable, CMOS integrated VCOs are often the first choice, fulfilling these requirements at mediocre performance. Combining VCOs with crystal or MEMS oscillators in a PLL revamps their rather modest long-term frequency stability and phase noise performance.

Crystal oscillators offer low phase noise, but are sensitive to the environment including pressure, temperature and humidity [136], [137]. During an MRI experiment, these effects do not undergo drastic changes. However, frequency stability may be hampered because of their sensitivity to magnetic fields and vibration [64], [138] caused by the fast switching gradients. Further, they are often contained in a magnetic housing which may disturb static magnetic field homogeneity, leading to image distortions.

Alternatively, MEMS oscillators generally have better shock and vibration performance due their lower mass [139], [140] and are widely available in non-magnetic cases, much smaller than comparable crystal devices. This comes at the cost of a higher output phase noise floor and close-in jitter.
Both types of oscillators undergo long-term frequency wander that may be prohibitive in MRI.

Besides influencing the oscillator itself, other scanner EMI effects can alter the frequency output of an oscillator including power supply modulation by Faraday induction or RF pulses. However, in [121] it was found that RF impact can be adequately addressed by proper shielding and input protection switching during excitation.

### 4.2.5 Clock correction

Oscillators are required for virtually any processing electronic device and in many cases accurate clocking is the basis for conclusive functionality. This is at risk if operated in-bore under scanner EMI.

To address the issue of in-bore modulation of the oscillator output, in the present work, we propose to simultaneously measure the phase deviations of the affected clock \( \tilde{\phi}_{\text{clk}} \) and use that information to time re-position digitized samples either in real-time or in a post processing step.

Precise knowledge of the timing modulation \( \tilde{\tau} \) of the sampling clock allows correction of acquired data \( \tilde{s} \) by adding an error compensating phase \( \phi_{\text{corr}} \).

\[
s[n] = \tilde{s}[n] \cdot e^{-i\omega_n \tilde{\tau}(nT)} = \tilde{s}[n] \cdot e^{-i\phi_{\text{corr}}[n]}
\]

and since \( \tilde{\tau}(nT) = \frac{\tilde{\phi}_{\text{clk}}[n]}{\omega_{\text{clk}}} \), it follows that

\[
\phi_{\text{corr}}[n] = \omega_n \tilde{\tau}(nT) = \frac{f_{\text{in}}}{f_{\text{clk}}} \cdot \tilde{\phi}_{\text{clk}}[n]. \tag{4.4}
\]

Before being added for correction, the compensating phase signal needs to be scaled by the ratio of input frequency to the actual sampling frequency. Both, the recorded error signal and the input signal have the same bandwidth.
4.3 Methods

4.3.1 Oscillators in-bore

To assess MRI EM environmental influence on clocking components, the impact of static and changing magnetic fields on a small sample of commercially off the shelf (COTS) oscillators was measured.

Perturbation by bench gradient simulation

Gradient influence on oscillators was assessed on four COTS devices on a custom bench gradient simulation and measurement setup. The used devices are listed in Table 4.1. Oscillators were mounted on a standardized PCB of size 2 cm x 3 cm with battery powered supply and additional voltage stabilization using a local ultra-low noise low dropout linear power regulator with high PSRR (LT3042, Linear Technologies). High PSRR suppresses gradient modulation on the supply. The four oscillators included two crystal and two MEMS devices with two types of outputs (LVDS) and CMOS and in both plastic and magnetic casings. Components were selected in the same price and performance range.

<table>
<thead>
<tr>
<th>Device</th>
<th>Component</th>
<th>Technology</th>
<th>Frequency [MHz]</th>
<th>Package</th>
<th>Output-type</th>
</tr>
</thead>
<tbody>
<tr>
<td>Crystal 1</td>
<td>SG-645</td>
<td>Crystal Oscillator</td>
<td>50</td>
<td>Plastic</td>
<td>CMOS</td>
</tr>
<tr>
<td>Crystal 2</td>
<td>KC5032</td>
<td>Crystal Oscillator</td>
<td>150</td>
<td>Metallic</td>
<td>CMOS</td>
</tr>
<tr>
<td>MEMS 1</td>
<td>ASVMP</td>
<td>MEMS</td>
<td>100</td>
<td>Plastic</td>
<td>LVDS</td>
</tr>
<tr>
<td>MEMS 2</td>
<td>ASMBD</td>
<td>MEMS</td>
<td>50</td>
<td>Plastic</td>
<td>CMOS</td>
</tr>
</tbody>
</table>

Table 4.1: Oscillator devices used for bench measurement and gradient field influence testing. Components were selected in the same price range and included MEMS and crystal devices.

The PCB with the mounted oscillators were placed sequentially on the gradient simulation setup. It consisted of a hand wound coil (48 turns, 6 layers, 1.2 mm wire spacing, 55 mm length) connected to a 600 W audio amplifier and a signal generator. With signal generator frequencies of
50 Hz-10 kHz and output currents of 50 mA-6 A, gradient slew rates from 50 mT s\(^{-1}\) to 120 T s\(^{-1}\) could be emulated. The distance of the coil to the PCB was fixed at 8 mm by a polymethylmethacrylat (PMMA) spacer and together with the PCB thickness (1.6 mm) and solder joints, the distance was 10 mm to the device unter test (DUT). Exciter coil and oscillator were centered by centering bolts.

Phase noise of the output was measured using a Spectrum Analyzer (NI-5603, National Instruments). Total jitter was calculated from the phase noise measurements over an integration range of 10 Hz to 1 MHz for all applied frequencies and currents. The modulation peak in the phase noise recording was measured at signal generator frequencies of 1 kHz, 3 kHz and 10 kHz over increasing modulation slew rate (set by the current). An averaged baseline measurement prior to modulation was acquired and subtracted for data analysis.
4.3.2 In-bore system perturbation

An in-bore receiver [121] was operated at 40 cm from the iso-center of the MR scanner (3 T & 7 T Achieva, Philips). It was connected via a bidirectional optical high-speed link to an out-field receiver station for communication and reference clock feed-in. The receiver was oriented such that the main PCBs were normal either to the X or Y gradient direction. Figure 4.3A&B shows the measurement setup.

Perturbations of acquired signals by gradient action was assessed by recording sine waves of 125 MHz-130 MHz at 3 T and 295 MHz-300 MHz at 7 T respectively (signal generators used: SMA100A & SMB100A, Rohde & Schwarz) at power of $-3$ dB below full range, while gradient echo, EPI and spiral sequences were run. In addition to $\partial B/\partial t$ and $B$ influencing the oscillator output frequency, Faraday induction can alter the tuning voltage of the employed VCO in the PLL.

To verify that predominantly the clock system is responsible for phase modulations experienced, additional bench measurements were performed. Using a hand wound exciter coil (see above) changing magnetic fields were locally induced on all components and sections on the processing board of the in-field system while recording a sine wave.
4.3. METHODS

Figure 4.3: System overview. (Continued on the next page)
Figure 4.3: (Previous page). A) shows the scanner bore with the in-field receiver system placed at 40 cm from the iso-center. There, gradients are becoming non-linear giving rise to more unpredictable changing magnetic fields. The sensitive clocking system on the main FPGA board is indicated in the red circle. B) shows the involved components of the out-field receiver system and the extension board with the hardware built to measure sampling clock modulations on the in-field system. C) schematic depiction of the signal flow of the proposed correction mechanism. 16 channels are sampled using a common encoding clock signal ($f_s$). This signal is derived from a PLL, locked to an external reference clock ($f_{ref}$) for multi-receiver synchronicity and long-term stabilization. The PLL has a field exposed crystal and a VCO. Changing magnetic fields cause disturbance of the PLL output which leads to time misplacements of acquired samples. One of the outputs of the PLL is fed back via the optical link ($f_{clk}$) and is acquired by custom build hardware. In an FPGA the phase of $f_{clk}$ is extracted and added to the 16 acquired data channels for correction.
4.3.3 Clock stabilization

With above measurements, clock modulation by gradients and its influence on the acquired signal were confirmed.

In cases where signals with high dynamic range need to be acquired, every mrad in stability is crucial. Examples include field probe measurements with sensitivities in the pT range [99] or neuroscience applications. The phase evolution of field probes contain the local dynamic field changes and empower gridded reconstruction for correction of magnetic field imperfections. The extraction of the field evolution necessitates the calculation of the derivative of the phase, amplifying any error in signal phase.

Therefore, we propose to correct this deterioration.

System overview

The proposed method for clock modulation correction is based on deploying a local (in-bore) low noise clock generator, phase locked to an external reference for long-term stability and multi-device synchronicity, as required for MR array imaging. Similar to existing techniques for the transfer of ultra-stable clocks via fiber networks [141], [142], erroneous phase modulations induced by reference clock spurs and channel perturbations (e.g. fiber noise and length changes) are compensated by a feedback mechanism.

Though, instead of the reference signal only, the effectively applied sampling clock signal is fed back. An out-field host system with dedicated hardware measures timing modulations of the sampling clock, which includes gradient modulation, the optical link added clock jitter, and clock reference imperfections. Comparison of the phase of the transmitted reference clock ($f_{\text{ref}}$) to the received sampling clock ($f_{\text{clk}}$) permits the extraction of a phase error signal ($\phi_{\text{clk}}$). As outlined above, knowledge of this error allows correction of acquired data. With proper FPGA processing this correction can be accomplished in real-time.

A system overview is given in Figure 4.3A-C.
Clock transmission

On the out-field system (Chapter 3), a 500 MHz reference clock ($f_{\text{ref}}$) is derived from an ultra-low phase noise OCXO (Abracon) dual PLL circuitry (LMK04828, Texas Instruments) with optional GPS locking. It is transmitted fiber-optically (PTX/PRX, Inneos) to the clock circuitry of the in-bore receiver. There, a PLL (Si5345, Silabs) locks an internal VCO to this reference and "cleans" the received clock from noise added by the optical uplink up to the programmed loop bandwidth [143].

The chosen uplink and reference frequency was 500 MHz, selected for three reasons: 1) it is a simple integer division to reach the desired 125 MHz for ADC clocking at 7 T and frequency division by N reduces the phase noise of the output signal [144] (expressed as $L(Nf) = 20 \log(N) \cdot L(f)$ in dB). 2) the optical uplink of the employed laser link has better jitter performance above 400 MHz. 3) if RF shielding of the PLL is insufficient and with a low input reference frequency, the phase detector in the PLL may experience additional phase transitions during the transmit pulse (Chapter 5 & [122]). Depending on filter bandwidth these cause the PLL to lose lock, introducing an unknown delay that spoils phase alignment of individual receive modules after each RF pulse. Having a reference clock and phase detector frequency higher than the Larmor frequency ensures in-time PLL re-locking or in case of a digital PLL distinct and known delay times.

An integer multiple ($f_{\text{clk}}$, 500 MHz) of the VCO derived ADC sampling clock ($f_s$, 125 MHz) is, together with the acquired data from 16 ADC channels, retransmitted back to the reference system via the multi-gigabit optical link. This signal was exposed to the same scanner EMI and Faraday induction with potential modulations and reference imperfections - up to the remaining differential trace (<2 cm in length at 0.9 mm spacing) from the clocking circuit to the ADC encoding input that was verified to be unaffected.
4.3. METHODS

Phase measurement

Various methods for measuring phase noise exist \cite{127}, \cite{145}, \cite{146}. On the out-field receiver system, a custom phase measurement hardware was built as plug-in extension module. The design employs a direct undersampling approach and allows for sub-picosecond phase measurement of the clock signal. The fed-back clock is band pass filtered (SXBP-507+, Mini-Circuits) to reduce noise aliasing, amplified (ADA4960-1) and digitized at 156.25 MSpS (LTC2107, Linear Technologies). The 500 MHz input signal is in the 7th Nyquist zone and gets aliased to 31.4 MHz. The ADC offers high SNR and low aperture jitter (45 fs RMS) at a nominal bit depth of 16 bit.

Digital processing

ADC data is transmitted in a high speed serial bus by LVDS. Digital processing is implemented in an FPGA (Z7045, Xilinx) and includes demodulation, decimation, filtering and extraction of the phase of the recorded clock signal. The digital filter chain consists of a computationally efficient CIC and a compensatory FIR filter. The CIC has a fixed order of four with programmable decimation factor. This allows the adjustment of the bandwidth of the clock phase signal to the bandwidth of the incoming data signals. The clock phase is extracted by a CORDIC (see Chapter 3). This phase is scaled according to (Eq 4.4) and time aligned to the incoming data stream using a multiplier and a buffer. Finally, it is added to the phase of the incoming data streams in real-time (Eq. 4.2.5). The extracted clock phase can optionally be stored as separate stream for offline calculations.

Clock stabilization verification

Results were obtained from measurements on 3 T and 7 T scanners (Achieva, Philips Healthcare, NL). The in-bore receiver was placed on the patient bed (4 cm from the iso-center, normal to the Y gradient) and a test signal was acquired on one channel (sine wave of 295 MHz...
at 7 T and sine wave of 130 MHz at 3 T, generated by an SMA100A Rhode&Schwarz). The clock correction was performed online in the FPGA fabric as well as offline for comparison.

Measurements with active clock correction and without gradient action were performed prior to enabling gradients to extract baseline stabilization performance. Similar to previous perturbation measurements, EPI and spiral sequences were played out along different gradient orientations. In addition to sequences, pulsed gradient tones \( (f_{\text{tone}}=10 \text{ Hz} - 10 \text{ kHz}) \) of 100 ms duration were produced within the limits of maximum available gradient slew rate \( (200 \text{ mT/m} \cdot \text{ms}^{-1}) \) and strength \( (40 \text{ mT/m}) \) in all gradient directions (X,Y,Z).

Besides recording test signals, EPI and spiral trajectories were monitored using a 16-channel field camera [56], connected to the in-bore receiver. For this experiment, the trajectories were first acquired with the receiver placed in front of the scanner bore (no gradient perturbation) representing the baseline reference. Subsequently, the trajectories were recorded with the receiver placed and operated in-bore.

The bandwidth of gradient amplifiers is limited to approximately 50 kHz. Therefore, field probe extracted k-space trajectories based on first order spherical harmonic expansion were filtered to 50 kHz using an FIR filter.

The influence of clock signal perturbations on an actual imaging experiment was simulated by encoding a phantom using the nominal trajectory (for both EPI and spirals) and decoding the data with the monitored trajectory, with and without clock correction. Likewise, coil data perturbation was simulated on a nominal trajectory with no phase error and by adding the phase error on a sine wave recorded in the worst case during an EPI sequence play-out.
4.4 Results

4.4.1 Oscillators in-bore

Perturbation by bench gradient simulation

Gradient bench simulation measurements showed that a changing magnetic field modulates the output of crystal resonators. It was found that the MEMS devices experience far lower to no relative modulation when compared to the tested crystal devices. Figure 4.5 shows the measurement results. In both crystal oscillators, a distinct modulation peak increased logarithmically with applied gradient slew rate. At a saturation point, crystals started to produce harmonics of their fundamental frequency. MEMS devices did not show clear modulation peaks or they were below the inherently high output phase noise. Crystal 1 experienced an increased phase noise level in the low frequency offset range once a changing field was applied.

Unfortunately, MEMS devices had an extensively higher total RMS jitter and phase noise floor. In most cases, the worst case modulations of the crystals were below the phase noise level of the MEMS devices at the given frequency. Figure 4.4 shows the RMS phase noise for the measured devices and the applied gradient slew rate. From the measured devices, only crystal 2 offered phase noise performance that lies in the required order of magnitude to be acceptable for a receiver system.

Due to limited acquisition and averaging time, the measurement uncertainty by the used phase noise analyzer is increasing with lower frequencies. Please note a missing data point in Figure 4.5 at the 10 kHz modulation with strongest current (high modulation slew rate).
Figure 4.4: Measured RMS jitter during externally applied changing magnetic fields on the four tested oscillator devices. MEMS devices exhibit a significantly higher total phase noise when compared to crystal devices. Crystal 2 showed an increase in RMS jitter with incrementing modulation slew rate. Except for crystal 2, all devices are unsuitable for acquisition of signals with mrad resolution required for field monitoring applications.
Figure 4.5: Results of the bench gradient simulation measurements on four oscillators. Crystal oscillators experience distinct modulation peaks when exposed to changing magnetic fields. MEMS devices lack such distinct modulation peaks, that however could be below the inherently high output phase noise.
4.4.2 In-bore system perturbation

Local field induction with the gradient bench measurement setup on the FPGA board showed that only the PLL with its supporting circuitry (VCO, Crystal) is prone to Faraday induction by changing magnetic fields, leading to a phase modulation of the recorded signal. Other sections of the board withstood strongest slew rates at no measurable impact on the signal, validating the suggested procedure of correcting the clock.

Further measurement results of in-bore operation are discussed and shown in Chapter 2.

4.4.3 Clock stabilization

Using the proposed clock stabilization hardware, the time-misplaced digitized data was compensated in both real-time and offline.

Hardware

Both lasers used for clock transmission of the bidirectional optical link operate at different wavelengths for wavelength division multiplexing on a single fiber and have differing maximum throughput rating (uplink 1.2 Gbit/s, downlink 3.5 Gbit/s). The uplink added an additional jitter of 250 fs to the reference clock, while the downlink only introduced 45 fs. For the proposed clock stabilization scheme this is beneficial, as it allows for more accurate corrections.

The clock phase measurement hardware exhibited a phase noise floor of 0.58 mrad RMS at 500 MHz input (equiv. 185 fs). This is enough to detect phase excursions of the in-bore ADC clock signal, whose clocking noise floor at the maximum dynamic range of 74 dB is the equivalent of 230 fs for 3 T operation and the additional jitter added by the optical uplink. Due to processing latency and run time differences, the clock measurement signal leads the data signals. This delay was 0.78 µs or
the equivalent of 121.8 system cycles and was accounted for in the developed FPGA core.

The difference of online correction in the FPGA and via post processing amounted to numerical precision (Figure 4.9).

**Correction with no gradient action**

Without active clock stabilization and no actively changing magnetic field, narrow band noise and spurs induced by the local clock limited the dynamic range and linearity of the in-field receiver (Figure 4.6). With active clock correction, the system experienced improved narrow band noise and reduced spur level. Phase SNR could be improved by 18 dB and SFDR by 16 dB. This indicates that clock deterioration in the optical uplink, the transmission electronics, and the in-field PLL generating the clock signal are not perfectly stable and profit from correction using the proposed feedback approach.

![Figure 4.6: Effect of the proposed clock correction scheme with no active external perturbation, i.e. no switching gradients. Shown here is the spectrum of the recorded signal phase with and without phase correction. The narrow-band phase noise of the acquired signal is improved and clock reference spurs are strongly reduced.](image-url)
Correction with gradient tones

Significant modulations of a recorded sine wave's phase was only experienced when gradient tones were played out along the gradient direction normal to the PCB (Y direction). No increase in amplitude of the phase evolution was observed for the other two directions (X, Z). This can be anticipated since a PCB typically has its largest area of potential Faraday induction normal to its surface. The result suggest that either the crystal or the VCO of the PLL has directional dependence or that other parts of the circuitry e.g. the PLL power supply, are influenced by gradients.

However, the gradient effect is well compensated by the proposed method. Any modulation of the phase after correction was within the noise floor. Equally, phase wander without gradient switching was improved. Figure 4.7 shows the measurement result and the correction effect on a demodulated signal with applied 100 ms, 500 Hz field tone at 7 T. This result affirms that the input signal is not modulated by gradient action and that the scaling by the frequency ratio is needed for correct compensation.
Figure 4.7: Correction effect of the clock compensation method on the phase of a recorded sine wave (295 MHz) in a 7 T scanner. The results showcase that with actively switched gradients, the sampling clock of the ADC is perturbed, which leads to a phase error in the recorded signal. Further, it becomes apparent that the input signal itself is not affected and that the modulation has directional dependence. This is explained by the PCB orientation in the field. Clock correction reduces the overall phase wander and corrects modulations to noise floor.
Correction on EPI and spiral sequences

Figure 4.8 shows the result of recording a 295 MHz sine wave signal in a 7 T scanner under an EPI sequence with full gradient duty cycle. Phase distortions of a maximum of 40 mrad peak-peak were generated in the recorded signal in the worst-case orientation.

The stabilization method reduced the phase modulation on the clocking signal from 40 mrad down to below 2 mrad in case of the EPI sequence, resulting in an RMS phase SNR improvement of 17.9 dB. In addition, using clock feedback the phase distortion was counteracted to below the noise floor of the receiver without clock correction. The noise floor was reduced to 1.2 mrad with active clock compensation turned on and hence frees up receiver dynamic range for imaging of up to 25 dB in phase.

Any residual modulation in the corrected signal is attributed to an actual modulation of the input signal during the EPI playout caused by limited common mode rejection in the higher frequency regime at such high dynamic range measurements. Vibrations during an EPI sequence have a higher bandwidth when compared to the single tone gradient switching case above. Even with fastest and strongest available gradient switching, the magnitude of a recorded signal changed only 10 ppm in the worst case. As with above gradient-tones, not all gradient directions affected the clock equally.

For illustration, the recorded clock phase error was used in a simulation to modulate $k_0$ of the nominal 4 interleave EPI trajectory. Such a clock phase error would give rise to corresponding EPI ghosting up to 1.4%.
Figure 4.8: Correction effect when running a 4-interleave EPI sequence (line colors) on a pure sine wave input tone of 295 MHz in the worst case orientation. Clock phase compensation leads to an overall decrease of phase noise floor and increases the dynamic range. The recorded phase modulation signal was used to perturb the nominal imaging trajectory ($k_0$) in simulation. An EPI ghosting of 1.4% can be expected.
Correction in field monitoring

ADC clock modulation affects all field probes equally and leads to a distortion of the $k_0$ term. Phase distortion of $k_0$ has the same effect as time re-allocating coil samples. The trajectory $k_0$ difference measured with and without clock compensation during a single-shot EPI read-out, repeatedly amounted to 10 mrad.

On the other hand, field probe SNR decreases with the decay of recorded FID, leading to a variance in the recorded trajectory. Besides probe SNR, fluctuations of the $k_0$ term stem from background field instability, and are associated with gradient heating and amplifier infidelity. Because the variance among the $k_0$ term of multiple shots of a monitored EPI trajectory out-bore was up to 45 mrad, a direct comparison of out-bore and in-bore monitoring with enabled clock correction is only meaningful if factors such as temperature were considered and ideally trajectories were monitored in- and out-bore in simultaneously.

The monitored single-shot EPI and spiral trajectories were used to simulate the influence on the decoding of a phantom. The corrected trajectory was used to encode the phantom while trajectories with and without correction were used for decoding. EPI ghosting amounted to 5 ppm while spiral blurring was 2.2 ppm in the case of 10 mrad $k_0$ disturbance.
Figure 4.9: Monitored single-shot EPI sequence from $^{19}$F field probes expanded to first order spherical harmonics ($k_0, k_{x,y,z}$). Clock phase modulation affects all recorded signals and therefore is visible in the $k_0$ term. A $k_0$ deviation of up to 10 mrad was corrected in real-time and offline. Likewise, a spiral sequence was monitored and the trajectory used to simulate the imaging process. Not correcting the clock phase led to a blurring of 2.2 ‰.
4.5 Discussion & Conclusion

With the current trend in MRI to operate sensitive receiver electronics in-bore, design considerations regarding the EMI environment are of great importance. Among others, the local clock of a receiver plays an important role in resolving high dynamic range coil or field probe signals.

In this work, it was re-stated that modulations or an increased phase noise level on the oscillator signal reduces dynamic range, regardless of receiver topology. Especially in case of a direct sampling architecture and for high input frequencies, poor clocking on the ADC is detrimental.

It was shown that gradient action modulates commonly used oscillator types and, on the example of a 16-channel in-bore receiver, it was demonstrated that for sequences with changing magnetic field during the read-out, image distortions occur.

Measurements suggest that MEMS devices exhibit lower field dependence when compared to crystal oscillators. In addition, more MEMS products are available in non-magnetic casings. However, due to their higher overall phase noise and the associated reduction in receiver dynamic range, they seem less suited as main timing device for broadband in-bore application.

Based on these insights, a method to counteract gradient induced clock modulations was presented. This is required to leverage the full performance of today's analog to digital converters in in-bore MRI receiver applications.

Several methods for ADC clock jitter compensation have previously been shown for other fields, ranging from CMOS jitter measurement circuits [147], to jitter estimation using pilot signals [148], and post-processing statistical methods [149]. However, these correction schemes only focus on actual clock timing jitter, i.e. phase noise of the encoding clock, and not on potential EMI signal modulations of the latter.

Meanwhile, the presented method effectively removes spurs and noise originating from imperfections of local clock sources and PLL circuits used on in-bore digitizers and corrects for scanner EMI induced signal distortions. It was shown that such a correction scheme boosts dynamic
range and compensates gradient modulations for significant improvements of phase SNR and SFDR of a receiver. Thereby, the advantage of full synchronicity provided by an up-link reference clock is combined with the robustness and accuracy of a down-link feedback.

With this correction scheme in place, in-bore sampling achieves state-of-the-art outfield receiver performance, even in presence of fast changing gradients.

Overall, it was confirmed that the in-bore receiver provides enough dynamic range to resolve limited field probe SNR of initial 40 dB [56] as discussed in Chapter 2. As a result of the meticulous design, gradient induced clock modulations led to image ghosting or blurring only in the %0-regime. For everyday scanning, this might not be detrimental. However, less stringently optimized receive electronics might experience higher distortions.

Phase stability is especially critical if one aims at ultra-high resolution field-monitoring applications [99] with in-bore digitized field-probes. There, the proposed clock stabilization scheme is vital.

Coherent acquisition of distributed receivers with low phase distortion is not only important in the field of MRI. The proposed method could be of interest for fields where receive electronics are exposed to disturbing EMI such as found mining applications.
Chapter 5

A Wearable MRI Receiver System

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• “A sub-1dB NF dual-channel on-coil CMOS receiver for magnetic resonance imaging”.

• “MR probe design with on-coil digital receiver”.

• “Integrated CMOS receiver for wearable coil arrays in MRI applications”.

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*Authors contributed equally to this work
5.1  Introduction

MRI faces fundamental limits in terms of sensitivity, speed and accuracy of the fields involved. One effective way of addressing the sensitivity limitation is simultaneous acquisition of data from multiple detectors, combined into RF coil arrays [15]. Apart from providing more signal than a body coil, they allow utilization of parallel imaging methods [17], [18], [150] to reduce total scan time, by exploiting data redundancy based on the different coil sensitivity distributions. More detectors with distinct sensitivities allow for higher imaging speeds. Hence, coil arrays with channel counts of 32 and more [20], [21], [31] are widely deployed to reach the image quality and acquisition speed required by modern clinical protocols. The demand for increasing the number of receive channels persists.

Employing tens of receive coils comes at the expense of as many cumbersome RF cables to carry the information from the coil to the digitizer out of the field. The size of these cables limit proper coil engineering and RF coupling among them can reduce SNR, and poses a potential safety risk due to occurring sheat currents. To tackle this, multi-channel receivers capable of high quality signal acquisition in the harsh EM environment of the MR scanner bore have lately been introduced [121], [151] (Chapter 2). They reduce the RF cable length to less than 1m and can be integrated into coils. Signals can be sent out of the bore by optical [32], [63], [121] or wireless links [57], [59], [60], [123].

However, to boost the number of receive coils in an array, bulky coaxial cables need to be eliminated altogether. For this, receivers need to be placed immediately on-coil, which is only possible with fully miniaturized electronics. Recently, the first receiver for MRI integrating the entire receive chain (i.e., from the LNA to filtering and digitization) was presented [122], [152] and renders on-coil digitization in the harsh electromagnetic environment feasible.

Despite all advances, the full potential of parallel imaging remains unexhausted by present-day coil arrays. Today, they are mostly rigid and built to fit a large population of patients. However, sensitivity is
only maximized if coils closely fit the target anatomy, hardly feasible with a fixed setup. Mechanical rigidity also prevents changes in posture, such as flexion of joints, and the array coils cage-like design is intimidating, especially for children.

To mitigate this, MRI detectors can be made flexible [153], [154] and even stretchable [47], [155], and can be produced with as simple processes as inkjet printing [156]. However, these developments have not yet found their way into day-to-day MRI, even if better sensitivity, enhanced patient comfort and workflow can be expected.

It turns out that combining these latest advancements into an actual wearable multichannel array presents itself as a non-trivial task. For wearability, size, weight and power restrictions of the electronics apply. A wearable array has to consider the great anatomical diversity of patients, which requires tuning and matching to be robust to changes in coil size and to effective coil loading, which varies from patient to patient and if wrongly implemented reduces net sensitivity.

Additionally, when scaling the number of detectors, the intricate interactions among coils and receivers have to be controlled - classical parameters for array design such as geometric decoupling are difficult to achieve with changing geometry.

Further, the immediate vicinity of the coils and the patient is a hostile as well as sensitive electromagnetic environment, which tolerates only the smallest of PCBs and virtually no magnetic material in its components.

On the way to radically change how MR signals are acquired, in this work, the feasibility of wearable, non-galvanic coil arrays with multiple distributed on-coil digital receivers is explored. First results based on a fixed dual-channel coil and an implementation of a stretchable array for wrist imaging, are presented.
5.2 Methods

5.2.1 The wearable MRI receive system - an overview

The following methods, components and PCBs were developed within the WearableMRI project, a collaboration of three institutes at ETH Zürich with the aim of enabling fully digital and wearable MRI detectors. The project was part of the Nano-Tera initiative\(^1\). Other than the development of the first fully integrated receiver frontend chip for MRI imaging [122], the project consisted of the elaboration of wearable communication links [157], the exploration of PCB design optimized for low EMI and low magnetic moment based on finite element method (FEM) simulations [124], advancements in stretchable coil design [155] and their automatic tuning [158], and a data recording and reconstruction framework [121], [159], [160]. Figure 5.1 shows an overview of the wearable system and the involved components.

Three basic building blocks make up the wearable on-coil frontend: the coil or the coil array, the tuning & matching board, and the digitizer-PCB hosting the custom ASIC. Each frontend connects up to two coils that receive the MRI signals. Digitized data is sent via an optical link to a receiver assembly placed out-field. There, signals of multiple on-coil receivers are collected, processed and image formation is performed. This out-field setup consists of a fiber optical receiver and clock distribution unit, a data collection hub and a reconstruction and control PC. Significant involvement in the development of the coil & coil interface, the clocking, the outfield setup and the reconstruction framework, as well as all MR measurements are part of this thesis.

In the following, these elements are detailed and the imaging subset of experiments, validating the concept of wearable in-field MRI receivers are presented.

\(^1\)www.nano-tera.ch Accessed: April 2018
Figure 5.1: System overview. A) Patient wearing a battery powered digital wrist coil array. Optical signaling, removes all galvanic connections to the scanner. B) The four channel stretchable wrist array coil consists of two optically connected dual-channel modules. C) Tune & match electronics attach coil and D) EMI optimized receiver PCB. E) Custom fiber receiver and clock distribution electronics interface four 2-channel wearable modules. For synchronization, each module receives a reference clock generated out-field. Boards are interfaced to an FPGA base board (*Chapter 2), for data synchronization and filtering. F) Outfield processing station collects data for image reconstruction on a PC.
5.2. METHODS

5.2.2 Receiver IC

The main component of the wearable MRI receiver is a custom ASIC, implementing the receiver chain needed for successful in-bore MRI image acquisition [122]. It employs a low-IF receiver architecture and features two receive channels with an input bandwidth of up to 450 MHz for scanner operation with static magnetic fields up to 10.5 T. It includes LNAs, frequency translation (passive mixers), filtering and digitization stages and was implemented in a 130 nm CMOS process, occupying an area of 22 mm².

It offers programmable gain (42 dB in 1 dB steps) to allow interfacing different coil array geometries and sizes, and can dynamically adjust the demodulation frequency, which allows recording of other nuclei such as ¹⁹F used in field-probes for concurrent trajectory monitoring. As opposed to traditional MRI preamplifier designs, the receiver’s LNA has a fully differential input for improved common-mode noise immunity. To enable decoupling of adjacent coils, the LNA input impedance needs to be highly mismatched to the source impedance of the coil, therefore, $Z_{\text{in}}$ can be tuned with its real part ranging from 25 Ω to 100 Ω. Analog-to-digital conversion is performed by means of a 3rd-order discrete-time delta-sigma modulator with more than 12-bit effective resolution.

The receiver IC is clocked by either a local, non-magnetic 26 MHz voltage controlled MEMS oscillator or a clock signal transmitted to the wearable system via a glass fiber from a stable out-field clock source. To provide low enough LO phase noise and the ability to lock to this external reference clock for multi-device synchronization, a digital PLL (DPLL) and a charge-pump PLL (CPPLL) are implemented (see synchronization section below).

An on-chip signal generator can be used to measure the resonance frequency of the attached coil and with the programmable input impedance, changes in coil loading can partially be compensated. The chip is programmed via a Serial Peripheral Interface (SPI) register interface.
5.2.3 Receiver PCB

A receiver board hosts the custom integrated receiver IC, where data is digitized, and interfaces it with the matching and detune circuitry of the coils on one side, and the glass fiber link to transmit the recorded data for further processing on the other side. The receiver board is a miniaturized and highly optimized PCB, shown in Figure 5.1D. It has 8-layers, a size of $2 \times 3\text{ cm}^2$ and a total thickness of $0.5\text{ mm}$ without components. The electronics for the wearable system need to withstand the harsh electro-magnetic environment inside the MRI bore during an active scan. Based on FEM simulations of the changing magnetic fields and the PCB, the board layout, geometry and circuit was optimized for operation in the MRI bore [124]. Only non-magnetic components where used and if not available, bare-die components were directly bonded onto the PCB.

On the board, digitized data of the receiver IC is fed via a parallel 1.8 V data bus to an auxiliary FPGA (Lattice, USA) where additional information such as channel number, I/Q labels, or synchronization commands are added. Data is transmitted as parallel bus of 16-bit at a rate of 80 MHz to a serializer (Texas Instruments, USA) where it is encoded with an 8b10b-code, before being transmitted at 1.6 Gbit/s via a glass fiber transmitter. The 80 MHz clock is generated by a non-magnetic MEMS oscillator (ASDMB, Abracon).

The FPGA communicates using an Universal Asynchronous Receiver Transmitter (UART) for receiver IC configuration and decodes additional up- and downlink commands and status bits such as the detune or malfunction signals, which are continuously being transmitted. Tuning and malfunction signals are made available on connector pins for use by the coil frontend.

The power supply of all components is sourced by a Lithium Ion battery (5000 mAh, 7.2 V) and level translated to the needed voltages by LDOs to minimize emitted EM noise and for better gradient immunity due to their high PSRR (see Figure 5.2).

This receiver PCB forms a general-purpose acquisition module.
5.2. METHODS

5.2.4 Coil interface

Coil specific circuitry for tuning, matching and detuning were implemented on a separate rigid-flex PCB with low-loss substrate (RO4350, Rogers, USA). It interfaces each coil to the fully differential preamplifier of the receiver IC through a Π-matching network embodying non-magnetic, high RF power, low equivalent series resistance (ESR) capacitors (American Technical Ceramics, USA) and high Q inductors (Coilcraft, USA) (Figure 5.2a).

The Π-matching network functions as a 90°-phase shifter for preamplifier decoupling[15] which relies on suppressing coil currents that lead to inductive coupling. Since the LNA offers a low impedance reflective input (500 Ω/50 Ω) using this transformation preamplifier coil decoupling can be established (about 14 dB by design). In addition, for noise optimal signal acquisition, the low impedance presented by the coil has to be transformed with low loss to the noise optimum input impedance (500 Ω) of the LNA. Due to its double-hump frequency response, a Π-matching approach offers robustness with respect to both subject-dependent changes in the resonant frequency of the coil and to variation in coil coupling induced by geometrical adjustment of the coil configuration [161].

Required space on the coil interface PCB was dominated by the detuning diodes and the large high voltage capacitors that have to sustain up to 300 V in detune state.

In addition to passive detuning by crossed high speed diodes, two methods for active coil Q-spoiling were implemented: 1) using actively biased non-magnetic PIN diodes (MACOM, USA) or 2) using depletion mode Gallium nitride (GaN) HEMT (CGHV27060MP, Wolfspeed, USA) which operate as junction field-effect transistor (JFET) switches, according to [162]. Active detuning using PIN diodes requires the wearable coil to be interfaced to the scanner's detuning voltages (+12 V/−5 V), which involves a remaining galvanic connection to the scanner. This deficiency can be mitigated by adopting the JFET Q-spoiling solution and embedding the detuning signal in the optical protocol. These transistors have to withstand the high RF powers and only few components providing
the needed specifications exist. An advantage of employing depletion mode transistors is that, due to their low resistance at a low gate-source voltage, they detune the coil per-default, rendering the coil operation safe, even in case the connection to the scanner is broken.

5.2.5 Coils

To demonstrate imaging capabilities of the wearable in-field receiver, two coil arrays were built, tested and applied for in-vivo imaging: 1) a fixed 2-channel dome-coil and 2) a 4-channel stretchable wrist coil. Besides, a fully flexible coil design based on liquid metal and used in conjunction with the wearable receiver, is described in [155]. A wearable knee coil is currently under development.

Detune circuitry of all coils were safety tested by surrounding mineral oil phantoms with the coil and acquiring images using the scanner’s body coil. Potential RF heating of tissue was excluded by exposing an agar gel phantom \((\varepsilon_r = 4.3)\) to 15-minutes of continuously played out RF pulses while tracking the gel temperature.

Several sources of loss can reduce image SNR and although the coil material contributes to them, the sample (i.e. the patient) usually dominates the total loss [163]. Therefore, a common way of coil characterization is the comparison of the coil’s quality factor \((Q)\) with and without sample loading \((Q_{\text{unloaded}}\) and \(Q_{\text{loaded}}\)). Where \(Q_{\text{unloaded}}\) contains information on the losses of the coil itself and \(Q_{\text{loaded}}\) the contribution from the sample. The higher the \(Q\), the better the SNR (less loss) [164].

The efficiency of a coil can be predicted by [165]:

\[
\text{SNR} = \text{SNR}_0 \sqrt{1 - \frac{Q_{\text{loaded}}}{Q_{\text{unloaded}}}},
\]

where \(\text{SNR}_0\) is the intrinsic SNR [165]–[167].
5.2. METHODS

Figure 5.2: The coils are separated from the patient by 3 mm spacer foam. Each coil is routed differentially to a PI matching network which performs a 90° phase shift for preamplifier decoupling [15] during receive. During transmit mode, a control voltage activates a PIN diode or a JFET for active detuning (for improved safety, also passive detuning is in place). The interface is realized on a rigid-flex PCB which conforms to the volunteer’s anatomy. The custom receiver ASIC contains novel, fully differential low-noise preamplifiers which allow strong common mode rejection. The receiver employs a low-IF receiver architecture and has a parallel data interface. Digitized data is encoded in a non-magnetic FPGA and after serialization is sent via an optical link to the out-field receiver system. Lithium ion batteries power the on-coil circuitry.
CHAPTER 5. WEARABLE MRI

Rigid 2-coil array

For proof-of-concept, a dual coil fixed array was constructed for receiver verification [122]. Two single-turn loop coils were made from adhesive copper tape and were mounted on a PMMA half-dome which allowed placement around a human wrist, the coil is shown in Figure 5.5A. Coil geometry of a single coil was 80 mm x 80 mm and conductors were capacitively split into sections to keep each section well below the RF wavelength in free space [167]. The coils were geometrically decoupled [15] and tuned for 3 T operation (127.7 MHz).

Wrist 4-coil array

For multi-receiver demonstration [168], a 4-channel flexible coil array for wrist imaging was designed following the approach of [47]. It is shown in Figure 5.3. A high-stretch bandage (IVF Hartmann AG, Switzerland) which allows stretching in one direction was used as the flexible and stretchable base material. A 5 mm copper braid (Easy Braid, USA) was chosen as coil conductor. Four coil elements were sewn onto a 690 mm stripe of the bandage. Each coil element has a size of 55 mm x 80 mm in the relaxed state (Figure 5.3A) extending to 89 mm x 80 mm when fully stretched. The location of the coils on the flexible substrate and their dimensions were selected to achieve geometric decoupling when in half-stretched state (Figure 5.3B). In addition, a 3 mm foam spacer was used to limit parasitic capacitive loading [47].
Figure 5.3: A) The wrist coil consists of four coils made from copper braid sewn onto a high stretch bandage. The placement and size of the coils is chosen to provide geometric decoupling and uniform coverage of the sample. B) The coil is wrapped around a phantom or a volunteer’s wrist and held in place using hook-and-loop fasteners sewn onto the stretch material. C) The stretchable wrist array allows for tight fitting around a volunteer’s wrist.

When the coil was placed on a phantom or a volunteer’s wrist (Figure 5.3C), tuning & matching was done manually using a network analyzer and decoupled pickup loops. The coils were tuned to 127.7 MHz and matched for use in a 3 T MRI system (Philips Ingenia, NL).

5.2.6 Receiver synchronization and scanner interface

For successful imaging, phase coherence among receive channels is crucial. In addition, RF transmit and receive electronics need to be synchronous, as varying trigger delays, random phase walks and frequency drifts lead to an accumulation of phase errors in the data and cause severe image artifacts. To synchronize asynchronous wearable receivers, they can individually be locked to the scanner’s main clock system by
wireless [169], or optical clock connections [121]. Here, locking of the on-chip PLL circuitry to an optically fed common external reference clock (208 MHz) was used.

However, for RF spoiling and receiver optimization, the excitation pulse phases on commercial MR systems are randomized by software in every cycle. These phase changes are vendor and sequence dependent and typically unknown. In addition, if insufficiently shielded, a local receiver PLL may lose synchronization due to the strong RF excitation pulses, with unreliable time to re-lock. Shielding typically leads to eddy currents that may be prohibitive in close proximity to the coil. Hence, additional means of synchronization have to be used. It has been demonstrated that sampling clocks can be synchronized by transmission of pilot tones during the receive period [170]. However, this requires undesirable additional hardware.

Therefore, instead of only acquiring imaging data, one can additionally record the excitation pulse ([79], [159]). Here, the fact that, due to limited coil detuning isolation during the B1+ phase, a strongly attenuated RF pulse signal is present at the digitizer inputs, is capitalized on. Even when losing lock at maximum incident RF power, PLL lock time is typically much faster than the RF pulse ring-down and significant parts of the transmitted RF can be recorded. Using the unwrapped pulse phase information every k-space line can be re-aligned. It has been shown that the variance of the estimates on the instantaneous signal phase using linear regression are identical to the Cramer-Rao bound if the signal was acquired with high enough SNR [79]. If the random phase cannot be extracted from the recorded pulse, the imaging sequence can be adjusted to allow the recording of the FID for a short period of time (few hundred nanoseconds) before gradient action starts.

5.2.7 Outfield processing and data reception

The receiver setup in the technical room is based on an in-bore FPGA base board (see Chapter 2) connected to an outfield processing station (see Chapter 3). Two add-on boards were developed [124] and interfaced to the FPGA base board: 1) a 4-module optical link transceiver
receives digitized MR signals from up to four in-bore wearable modules (total of 8 channels) by optical fiber modules (AFBR, Avago) and interleaves data on a local flash-FPGA (MachXO03LF, Lattice) for forwarding to the base board. 2) a 4-module optical clock reference transmitter board where an external reference clock (208 MHz, SMA100A, Rohde&Schwarz) is split and fed to an electro-optical converter (Finisar) for fiber transmission. Figure 5.1E shows the assembled hardware.

On the FPGA base board (FPGA: Artix-7, Xilinx) the incoming data stream is decimated by a factor of 8 using a CIC - from the incoming rate of 7.68 MHz to 960 kHz bandwidth - and passed through an FIR filter (pass-band = 800 kHz, stop-band = 960 kHz, ripple = 0.1 dB, stop-attenuation = 60 dB).

Signals are forwarded through a multi-gigabit optical link to the outfield processing computer where data is stored and image reconstruction is performed using Matlab (The MathWorks, Natick, USA) (Figure 5.1F). Image formation of the recorded gradient echo sequences employs phase extraction of the recorded RF pulse with correction of k-space data, a 2D tapered cosine k-space filter and FFT. Single coil images are combined using the sum of squares method.

5.2.8 Imaging experiments

With the dual-channel dome coil array, an image of a wrist of a healthy volunteer was acquired on a 3 T human scanner (Achieva, Philips, Netherlands). A standard gradient echo sequence with imaging parameters according to [47] was selected. Ten averages (NSA) and the following scan parameters were used: TR = 132 ms, TE = 8.3 ms, flip angle = 30°, pixel bandwidth = 217 Hz and a voxel size = 0.2 x 0.2 x 4 mm$^3$. The resulting image is shown in Figure 5.5B.

The 4-channel wrist array employing two wearable modules, was used to acquire a phantom image on a 3 T system (Ingenia, Philips, Netherlands) using a fast field echo (FFE) sequence (TR = 230 ms, TE = 8 ms, flip angle = 70°, NSA = 5, voxel size = 0.45 x 0.45 x 5 mm$^3$). The
phantom fluid was CuSO₄ doped saline water to emulate human tissue [171].

In-vivo images (Multi-slice FFE, 10 slices, TR = 132 ms, TE = 8.3 ms, flip angle = 30°, NSA = 8, voxel size = 0.45 x 0.45 x 5 mm³) of two healthy volunteer’s wrists of different sizes, requiring the stretching of the wrist-coil, were acquired.

*Figure 5.4: The phantom image acquired with the 4-channel wrist array shows uniform coverage and no signal voids, which indicates good detuning, decoupling and no B₀ distortions due to magnetic electronic components which are located in close proximity to the coil*
Figure 5.5: A) 2-channel coil demonstrator with fixed coil geometry for human wrist imaging. Detail view of the receiver PCB with its components. Volunteer’s wrist in place for imaging procedure. B) One slice from a human wrist imaged using the setup at a 3 T scanner. TR = 132 ms, TE = 8.3 ms, flip-angle = 30°, pixel-bandwidth = 217 Hz, voxel size = 0.2 x 0.2 x 4 mm³ ©IEEE 2017

1 published in [157], 2 published in [122], [152], 3 published in [124]
5.3 Results

5.3.1 Receiver IC

The receiver IC achieved sub-1 dB noise figure and a dynamic range of up to 81.9 dB at 1 MHz output bandwidth. Power consumption of the chip was 240 mW per channel. Maximum input power $P_{\text{inmax}}$ was $-20$ dBm, third-order input intercept point (IIP3) was $0$ dBm, and the 0.1 dB compression point was at $-23$ dBm. Under changing magnetic fields with an EPI running, the total LO rms jitter was 297 ps over an integration range of 1 Hz to 500 kHz and dropped to 3.3 ps with activated on-chip DPLL, reflecting an improvement of two orders of magnitude, and enough for on-coil high dynamic range imaging. Comprehensive results of the receiver IC measurements are given in [122].

5.3.2 Receiver module

The total power consumption per channel of an assembled on-coil receiver module was 1.05 W with most power being consumed by the serializer and optical module. Some residual magnetic components caused susceptibility artifacts when imaging a slice very close to the electronics. Since the receiver employs a low-IF architecture, resulting in potentially high DC offsets, the demodulation frequency has to be chosen such that the imaging bandwidth is separated several kHz away. Demodulation frequency ($f_{\text{demod}}$) selection can have a strong impact on spurs occurring from aliased harmonics originating the local oscillators (80 MHz, 26 MHz). By appropriate selection of $f_{\text{demod}}$ and offline demodulation for image alignment, frequency spurs in the imaging band were below $-80$ dB. Comprehensive results of receiver module measurements are given in [124].
5.3.3 Coil interface

The performed system and patient safety evaluation tests confirmed very good detune isolation by not distorting the field of the MR scanner’s whole body transmitter and receiver and with a maximum single ended input swing of 1.8 V also meeting the LNA tolerances. Temperature measurements showed a relative heating below 1.5°C over the total exposition time.

5.3.4 Single receiver 2-coil imaging

Figure 5.5B shows the acquired image of the wrist of a human volunteer.

5.3.5 Wrist 4-coil array imaging

Tuned to 127.7 MHz the coils provided a $Q_{\text{loaded}}/Q_{\text{unloaded}} = 24/85$. Only marginally degrading intrinsic SNR. Phantom imaging results are shown in Figure 5.4). The in-vivo images of a human wrist show good coverage and image uniformity (Figure 5.7). Stretching by 18% shifted the center frequency by 3.2 MHz which was compensated on a per subject basis. Stretching of the wearable wrist coil did not degrade image quality (Figure 5.6).
Figure 5.6: Two slices (out of 10) from imaging two healthy volunteer’s wrists of different sizes. (FFE, TR = 132 ms, TE = 8.3 ms, flip angle = 30°, NSA = 4, voxel size = 0.25 x 0.25 x 5 mm$^3$). The stretchable coil seamlessly adjusted to the different wrist diameters without apparent image degradation.
5.3. RESULTS

Figure 5.7: In-vivo images of a healthy volunteer, acquired using a standard gradient echo sequence (FFE) (\(TR = 132 \text{ ms}, \ TE = 8.3 \text{ ms}, \ flip \ angle = 30^\circ, \ NSA = 4, \ voxel \ size = 0.25 \times 0.25 \times 5 \text{ mm}^3\))
5.4 Discussion

Several advantages arise from wearable coil arrays for MRI. Arguably, the most important is patient experience. Wearables help mitigating claustrophobia and render the scanning situation less intimidating, especially for children. This reduces the need for re-scans, which are uncomfortable and costly. Flexible coils also improve diagnostic quality by enabling new insights with the ability to depict joints in motion [47]. Moreover, by bringing the coils as close as possible to the target anatomy, imaging performance can be strongly improved for higher resolution images, or reduced scan time. Another important aspect is workflow. Today, patient setup time takes up a significant amount of each MRI examination, proliferating cost. If patients could wear the coils like a piece of clothing and prepare for the examination by correctly placing the coil in a preparation room, time of occupying the scanner can be reduced.

The presented work reflects the current status on the way of transforming MRI receive array's into entirely wearable devices. A wearable receiver system comprising of a stretchable multi-channel coil array with fully integrated custom receivers for RF signal detection directly on-coil has been implemented. It was demonstrated that on-body and in-field data reception of multiple wearable devices during an MRI scan is possible. This is owed to integration and careful optimization of involved components. The platform enables lighter, safer and cheaper receiver arrays with higher channel counts.

With the custom design of the key integrated circuits, small form factor on-coil receivers can be built. The realized fully differential analogue design with a high degree of common-mode rejection, make supplies, ground and signals less vulnerable to aggressive distortions from gradients and drastically simplifies coil design by removing the need for baluns. Compared to HBT/HEMT processes typically employed for high fidelity RF applications, the standard CMOS process of 130 nm offers a low volume threshold and comparably low production cost, which is required to meet the quantities for MRI applications. Nevertheless, sufficient analogue signal performance is obtained for digitizing the RF
5.4. DISCUSSION

Signals of receiver arrays directly on each coil element. The receiver features sub-1dB noise figure and allows operation on most currently installed scanners (1.5-10.5 T).

Miniaturization and optimization of electronic components and PCBs allow electronics to be placed close to the volume being imaged. However, even with available high-performance clusters, board design optimization by FEM is limited by the circuit complexity and showed that not all eddy currents can be removed. Nevertheless, placement and eddy current simulation can significantly help reducing artifacts in images.

Sending received signals via optical fibers from within the scanner bore to the reconstruction PC is a cheap, flexible and an EMI insensitive solution. Traditional cabling struggles are relieved and signal integrity and safety issues superseded.

The system was successfully applied to invivo imaging, presenting the worlds first reported image created by an on-coil digitized wearable multi-channel array. Yet, it needs fine tuning. One of the remaining challenges is the relatively high power consumption and heat generation, in close vicinity of the patient. An effective approach to achieve this is to replace the power hungry serializer chip or even implement the data serialization for optical transmission directly on the receiver IC. Instead of batteries, power could be short term harvested from gradients[172], optically transmitted [124] or even received wirelessly [173].

In general, the system would benefit from further integration: integrating the power supply LDOs removes the package causing susceptibility artifacts, that otherwise need to be directly bonded as bare die on the substrate. Integrating the drivers needed for optical signaling helps reducing the overall form factor, and a digitally programmable capacitor array could remove discrete varicaps for integrated tuning & coil matching. Integration can also include further signal processing steps to reduce the data rate for efficient data connections. Remaining space is required by the detuning circuitry (diodes or JFET) and could be reduced by employing high RF power sustaining MEMS switches [174].

Currently, the fiber optic link is impacted by excitation RF, disrupting communication and clocking during the pulse. This was mitigated by
post alignment, should however be solved by shielding the affected traces and by using higher clocking frequencies for faster PLL re-locking.

When scaling the number of coils, coupling among the coils is a challenge. This is especially true with stretchable coil arrays that face varying coil coupling depending on the patient anatomy. Besides improving the LNA input impedance for better preamplifier decoupling, the system can be complemented by adaptive electronics that sense effective port impedances and automatically adjust variable matching networks to compensate for load and coil geometry changes [175]. This empowers optimal SNR yield. For the current receiver IC, frequency sensing has been shown by [158].

Future applications are not limited to imaging only. By expanding the receiver chip with a power amplifier, the internal signal generator can be used to excite and acquire data from NMR field probes. NMR field probes have been reported to give new physiological insights by recording subtle field fluctuations related to physiological processes inside the body [99], [104] and can be used for bulk motion tracking during MRI scans [45], [46].
Chapter 6

Conclusions and Outlook

This work addressed some of the challenges faced by present-day MRI acquisition.

A receiver was developed that stands out in that it operates in the harsh environment of the MR scanner bore, while matching the performance of traditional out-field spectrometers and even allows sensitive field-monitoring.

Clocking of in-bore devices was thoroughly studied in general and a solution was presented to boost receiver accuracy under gradient operation. Clock feedback helped stabilizing sampling under changing magnetic fields and improved overall dynamic range.

The modular spectrometer covers a wide frequency range to serve MRI devices with a static magnetic field of 1 T to 9.4 T, and supports seamless acquisition of a large range of signals thanks to a broadband input. This receive flexibility has many applications. For example, it permits simultaneous recording of field probe and coil data or the acquisition of X-nuclei with dual tuned coils on a single system.

Several novel applications that have not been possible with previously available systems were demonstrated. This includes the full synchronous acquisition of imaging and field monitoring data from multiple receivers
on a common time base to allow smooth reconstruction of demanding sequences, which to this date required tedious post alignment.

In-bore units reduce both hardware cost and cabling efforts. This is particularly beneficial when integrating and combining a receive coil array which in addition to imaging coils, includes field monitoring probes for precise trajectory measurement and sensors for prospective motion correction or vital sign observation. Such an "all-inclusive" coil benefits from the optical connection and the synchronous acquisition and processing of both imaging and sensor data. Together with available nimble reconstruction methods, many of the remaining scanner imperfections can be circumvented in a single device, increasing MR image quality.

The sheer amount of data produced by large scale coil arrays and emerging auxiliary sensors is difficult to handle. In this work a solution was presented and digitization and processing challenges were addressed, to enable both, enhanced image reconstruction based on sensor information, and run-time updates of scanner operation. For this, a versatile acquisition and processing platform was built. In combination with the in-field receiver, this makes up for a complete MRI spectrometer that is truly extensible and, compared to existing systems, not duty cycle limited. With the proposed transition from a rigid receiver design to a distributed, reconfigurable acquisition philosophy, where coil data is complemented by substantial auxiliary data inputs in real-time, novel possibilities for MR methodology emerged.

The usefulness of having a versatile and programmable digital processing pipeline was demonstrated by enabling multi-rate acquisition, which was applied to acquire data with minimal dead-time in ZTE-imaging at only minor increase of data [71]. Further, the power of simultaneous real-time k-space trajectory extraction and coil compression in hardware was illustrated, which resulted in significant data savings without loss of flexibility and image quality. One can envision real-time data pre-processing to be combined with rapid cloud reconstruction [176] to permit unprecedented interventional imaging possibilities. Dynamic imaging and recent machine learning based approaches that exploit spatiotemporal data redundancy might equally profit from agile parallel processing power.
The novel out-field receiver provides the basis for future development and surpasses the abilities of the existing platforms in a compact form factor. Its modularity, the high connectivity options and the reconfigurable, programmable digital subsystem allows fast and comparably easy implementation of future advanced applications and integration of other sensor and actuator front-ends; for instance a gradient controller or an RF transmit interface. Two evident applications are real-time patient motion detection with rapid sequence updates and scanner field stabilization. Both benefit from the available hardware based processing with low-latency. Hence, carrying on the started work, a gradient and shim interface combined with an appropriate controller allows to close the loop [39], [46] even faster, and thanks to the ability of autonomous operation would render the system a desirable plug-in and stand-alone product. Real-time slice-wise dynamic shim updates are another promising application. Further, full console integration could be envisioned to for example drive RF excitation based on B1+ monitored data.

Finally, the presented integration of a custom MRI receiver chip is an essential step towards replacing the cage-like detectors used in today’s clinical MRI for improved patient comfort, sensitivity and new insights enabled by the ability to easily change patient posture.

Few challenges remain, including efficient power supply for reduced heat generation, coil decoupling and automatic tuning and matching as well as miniaturization of T/R switching. Future versions of the receiver chip should include an RF generator for field probe excitation and all the necessary driver circuitry to directly interface an optical data link.

With today’s possibilities of circuit integration and the advances in flexible coil array technology, it might be that future MRI coil arrays will be wearable textiles containing adaptive detector arrays with miniaturized on-coil receivers. All coupled with reduced connectivity for ultimate patient comfort and sensitivity.

To conclude, we addressed MRI acquisition and processing hardware and went from out-field to in-bore acquisition, from offline to large scale real-time processing and even introduced on-body digitization.
Appendix A

Further contributions

A considerable amount of (engineering-)work accomplished during this PhD remains largely undocumented here and is partially documented elsewhere. With this work several people from the IBT and other groups could advance their research, including e.g. gradient impulse response function (GIRF) prediction from running sequences [177], [178], some significant insights into GIRF temperature behavior and their prediction for model adjustments [179], [180], a new type of stretchable array coil based on liquid metal [155], a novel method of shimming [181], [182], high-speed T/R switching [183], gradient temperature supervision in ultra-short T2* imaging and fMRI studies [184], [185], or the control soft- and hardware for excitation pulses for a new gradient system [12], [186], [187]. Besides, all scanners used by IBT were equipped with a highly precise triggering system that emerged from this work and several other hardware tools were built. It is rewarding to see one's work being appreciated in practice.

Various side projects benefited from the experience gained in the above described work. This includes a wearable vital sign monitoring device capable of working in-bore and connected via bluetooth low energy (BLE) [188]. We showed that BLE can be used for reliable communication during a scan, making today's often needed and time consuming exchange of monitoring devices between out-field and in-field
operation obsolete. Wireless technology will be a key enabler for future coil and monitoring devices.

In another project (manuscript in preparation), the feasibility of increasing power supply efficiency and reducing related heat generation was assessed. For this, we compared power supply topologies and their applicability for in-bore electronics [124]. Buck-converter, switched capacitor, and LDO devices were compared in terms of scanner EMI and efficiency. While buck-converters are very noisy, measurements suggest that shielded switched-capacitor regulators could potentially replace currently used LDO as in-bore supply regulators with much higher efficiency.
# Acronyms and Symbols

## Acronyms

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Description</th>
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<tbody>
<tr>
<td>$^{19}$F</td>
<td>Fluorine-19</td>
</tr>
<tr>
<td>$^1$H</td>
<td>hydrogen</td>
</tr>
<tr>
<td>2.5D</td>
<td>2.5-dimensional</td>
</tr>
<tr>
<td>ADC</td>
<td>analog-to-digital converter</td>
</tr>
<tr>
<td>ASIC</td>
<td>application specific integrated circuit</td>
</tr>
<tr>
<td>$B_0$</td>
<td>static magnetic field</td>
</tr>
<tr>
<td>$B_1$</td>
<td>radio frequency field</td>
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<tr>
<td>BGA</td>
<td>ball grid array</td>
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<tr>
<td>BLE</td>
<td>bluetooth low energy</td>
</tr>
<tr>
<td>BW</td>
<td>bandwidth</td>
</tr>
<tr>
<td>CIC</td>
<td>cascaded-integrator-comb</td>
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<tr>
<td>CMOS</td>
<td>complementary metal-oxide semiconductors</td>
</tr>
<tr>
<td>CORDIC</td>
<td>COordinate Rotation Digital Computer</td>
</tr>
<tr>
<td>COTS</td>
<td>commercially off the shelf</td>
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<tr>
<td>Acronym</td>
<td>Definition</td>
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<tr>
<td>CSF</td>
<td>cerebrospinal fluid</td>
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<tr>
<td>DC</td>
<td>direct current</td>
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<tr>
<td>DDC</td>
<td>digital-down-conversion</td>
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<tr>
<td>DDS</td>
<td>direct digital synthesis</td>
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<tr>
<td>DNL</td>
<td>dynamic nonlinearity</td>
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<tr>
<td>DUT</td>
<td>device under test</td>
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<tr>
<td>EM</td>
<td>electromagnetic</td>
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<tr>
<td>EMF</td>
<td>electromotive force</td>
</tr>
<tr>
<td>EMI</td>
<td>electromagnetic interference</td>
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<tr>
<td>ENOB</td>
<td>effective number of bits</td>
</tr>
<tr>
<td>ENR</td>
<td>equivalent noise resistance</td>
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<td>EPI</td>
<td>echo-planar imaging</td>
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<tr>
<td>ESR</td>
<td>equivalent series resistance</td>
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<td>FEM</td>
<td>finite element method</td>
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<tr>
<td>FFE</td>
<td>fast field echo</td>
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<tr>
<td>FFT</td>
<td>fast fourier transform</td>
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<tr>
<td>FID</td>
<td>free-induction-decay</td>
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<tr>
<td>FIFO</td>
<td>first-in-first-out</td>
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<tr>
<td>FIR</td>
<td>finite impulse response</td>
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<tr>
<td>fMRI</td>
<td>functional magnetic resonance imaging</td>
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<tr>
<td>FOV</td>
<td>field of view</td>
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<tr>
<td>FPGA</td>
<td>field-programmable gate array</td>
</tr>
<tr>
<td>GaAs</td>
<td>Gallium-Arsenide</td>
</tr>
<tr>
<td>GaN</td>
<td>Gallium nitride</td>
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<tr>
<td>GIRF</td>
<td>gradient impulse response function</td>
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<tr>
<td>GPS</td>
<td>global positioning system</td>
</tr>
<tr>
<td>GPSDO</td>
<td>GPS disciplined oscillator</td>
</tr>
<tr>
<td>HEMT</td>
<td>high-electron-mobility transistor</td>
</tr>
<tr>
<td>Acronyms</td>
<td>Description</td>
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<td>-----------</td>
<td>--------------------------------------------------</td>
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<tr>
<td>HSDL</td>
<td>high-speed data link</td>
</tr>
<tr>
<td>IC</td>
<td>integrated circuit</td>
</tr>
<tr>
<td>IIP3</td>
<td>third-order input intercept point</td>
</tr>
<tr>
<td>IO</td>
<td>input-output</td>
</tr>
<tr>
<td>JFET</td>
<td>junction field-effect transistor</td>
</tr>
<tr>
<td>LDO</td>
<td>low-dropout regulator</td>
</tr>
<tr>
<td>LNA</td>
<td>low-noise amplifier</td>
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<tr>
<td>LO</td>
<td>local oscillator</td>
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<tr>
<td>LVDS</td>
<td>low voltage differential signaling</td>
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<tr>
<td>MAC</td>
<td>multiply-accumulate</td>
</tr>
<tr>
<td>MCX</td>
<td>micro-coaxial</td>
</tr>
<tr>
<td>MEMS</td>
<td>microelectromechanical systems</td>
</tr>
<tr>
<td>MHz</td>
<td>megahertz</td>
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<tr>
<td>MIMO</td>
<td>multiple-input-multiple-output</td>
</tr>
<tr>
<td>MP-RAGE</td>
<td>magnetization prepared rapid gradient echo</td>
</tr>
<tr>
<td>MR</td>
<td>magnetic resonance</td>
</tr>
<tr>
<td>MRI</td>
<td>magnetic resonance imaging</td>
</tr>
<tr>
<td>MRT</td>
<td>Magnetresonanztomographie</td>
</tr>
<tr>
<td>MSB</td>
<td>most significant bit</td>
</tr>
<tr>
<td>NF</td>
<td>noise figure</td>
</tr>
<tr>
<td>NMR</td>
<td>nuclear magnetic resonance</td>
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<tr>
<td>OCXO</td>
<td>oven-controlled crystal oscillator</td>
</tr>
<tr>
<td>PCB</td>
<td>printed circuit board</td>
</tr>
<tr>
<td>PLL</td>
<td>phased locked loop</td>
</tr>
<tr>
<td>PMMA</td>
<td>polymethylmethacrylat</td>
</tr>
<tr>
<td>Acronym</td>
<td>Definition</td>
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<tr>
<td>PNS</td>
<td>peripheral nerve stimulation</td>
</tr>
<tr>
<td>PSRR</td>
<td>power supply rejection ratio</td>
</tr>
<tr>
<td>RF</td>
<td>radio frequency</td>
</tr>
<tr>
<td>RMS</td>
<td>root mean square</td>
</tr>
<tr>
<td>ROI</td>
<td>regions of interest</td>
</tr>
<tr>
<td>SAR</td>
<td>specific absorption rate</td>
</tr>
<tr>
<td>SATA</td>
<td>Serial Advanced Technology Attachment</td>
</tr>
<tr>
<td>SFDR</td>
<td>spur free dynamic range</td>
</tr>
<tr>
<td>SNR</td>
<td>signal-to-noise ratio</td>
</tr>
<tr>
<td>SoC</td>
<td>System-on-Chip</td>
</tr>
<tr>
<td>SPI</td>
<td>Serial Peripheral Interface</td>
</tr>
<tr>
<td>$T_1$</td>
<td>longitudinal relaxation time</td>
</tr>
<tr>
<td>$T_2^*$</td>
<td>transversal relaxation time</td>
</tr>
<tr>
<td>TCXO</td>
<td>temperature-compensated crystal oscillator</td>
</tr>
<tr>
<td>TE</td>
<td>echo time</td>
</tr>
<tr>
<td>TFE</td>
<td>turbo field echo</td>
</tr>
<tr>
<td>T/R</td>
<td>transmit-receive</td>
</tr>
<tr>
<td>TR</td>
<td>repetition time</td>
</tr>
<tr>
<td>TTL</td>
<td>transistor-transistor logic</td>
</tr>
<tr>
<td>UART</td>
<td>Universal Asynchronous Receiver Transmitter</td>
</tr>
<tr>
<td>VCO</td>
<td>voltage-controlled oscillator</td>
</tr>
<tr>
<td>VCXO</td>
<td>voltage-controlled crystal oscillator</td>
</tr>
<tr>
<td>ZTE</td>
<td>zero-echo-time</td>
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References


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REFERENCES


[174] \textit{Menlo Micro}.


List of Publications

Journal Articles


List of Publications


Conference Contributions


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