A Magnetic Field-to-Digital Converter Employing a Spin-Torque Nano-Oscillator

Dagur I. Albertsson, Johan Åkerman, and Ana Rusu

Abstract—In this work, a novel magnetic field-to-digital converter based on emerging spin-torque nano-oscillators (STNOs) is proposed. The architecture is inspired by voltage controlled oscillator (VCO)-based analog-to-digital converters (ADCs) which have shown inherent first-order noise shaping of both quantization- and phase-noise without the need for feedback. In the proposed architecture, the STNO acts both as a magnetic field sensor and VCO. The architecture's performance is evaluated in terms of signal-to-noise and distortion ratio (SNDR) utilizing Verilog-AMS modeling, where a macrospin model fitted to experimental data is employed for accurate description of the STNO operation. The presented simulation results demonstrate the potential of the STNO-based magnetic field-to-digital converter architecture.

Index Terms—Spin-torque Nano-Oscillator, Spintronics, Verilog, Magnetic field sensor.

I. INTRODUCTION

OVER the last decades, sensor based systems have revolutionized many industries. Specifically, magnetic field sensors are used in magnetic storage devices [1], biosensing applications [2], structural health monitoring [3], automobiles and a wide range of applications in factories [4]. The most common magnetic field sensors employed in these applications include Hall-effect, anisotropic-, giant- and tunneling-magnetoresistance sensors [5]. Alternatively, a relatively new type of magnetic field sensors that might allow for higher spatial resolution are based on spin-torque nano-oscillators (STNOs) [6], a novel oscillator which has attracted significant attention in recent years. Their nano-scale size along with wide frequency tunability make STNOs an interesting device for a number of applications, e.g. wireless communication systems [7], neuromorphic/non-boolean computing [8], [9] and as magnetic storage read/write probes [10], [11] or magnetic field sensors [6]. However, the STNOs current performance in terms of output power and phase noise limits their applications. The STNO applications often rely on monitoring the operating frequency which brings challenges for the analog blocks in the readout where wide bandwidth, low noise and high gain are needed. Specifically, STNO magnetic field sensors are based on tracking changes in the operating frequency, which can be multi-GHz for small variations in magnetic

Fig. 1. (a) Illustration of the MTJ STNO and (b) Graphical representation of terms in the LLGS equation.

fields. Consequently, STNOs have the potential of being used as high resolution magnetic field sensors. A typical readout circuitry for STNO-based magnetic field sensors proposed in [12] requires a frequency-to-voltage converter followed by an analog-to-digital converter (ADC), which are challenging to design and consume substantial area and power. In this work, we take a step further by proposing a potentially more power and area efficient readout architecture, which is a STNO-based magnetic field-to-digital converter. The proposed architecture is inspired by voltage controlled oscillator (VCO)-based ADCs, which are mostly digital CMOS circuits that bring the advantages of low power, small area and technology scaling [13], [14].

The paper is organized as follows. Section II describes the fundamental characteristics of STNO devices and the proposed architecture. Section III covers simulation results and discusses possible performance enhancements of the magnetic field-to-digital converter. Finally, section IV draws conclusions.

II. PROPOSED MAGNETIC FIELD-TO-DIGITAL CONVERTER

A. STNO operation

Spintronic oscillators are nanoscaled, current controlled high frequency devices which can be categorized in terms of material structure and/or geometry [15]. The most common STNOs are either based on magnetic tunnel junctions (MTJs) or spin-valves. MTJs are made of an insulator spacer, such as MgO, sandwiched between two ferromagnetic layers (typically CoFeB or NiFe), as it is shown in Fig. 1.(a), while in spin-valves the insulator is replaced by a non-magnetic metal spacer such as Cu. These devices feature different characteristics where MTJ based STNOs have shown higher output power while spin-valve oscillators have a wider frequency range [15].

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Due to the higher output power of MTJ STNOs, they have gained more interest in recent years [15]. The MTJ STNO is composed of a pinned ferromagnetic layer (PL) designed to have a constant magnetization direction while the free layer (FL) can precess when a magnetic field, H_{ext} and a DC current, I_{DC} are applied to the device. The precession is explained in terms of the Landau-Lifshitz-Gilbert equation with a Slonczewski spin-transfer (LLGS) term as [16], [17]:

$$\frac{d\mathbf{M}_{FL}}{dt} = -\gamma[\mathbf{M}_{FL} \times \mathbf{H}_{eff}] + \mathbf{T}_\alpha + \mathbf{T}_S \quad (1)$$

where $\mathbf{M}_{FL}$ is the FL magnetization direction, γ is the gyromagnetic ratio, $\mathbf{H}_{eff}$ is the effective magnetic field inside the FL, $\mathbf{T}_\alpha$ is the magnetization intrinsic damping torque and $\mathbf{T}_S$ is an antidamping torque as shown in Fig. 1.(b). When a DC current, I_{DC} , is passed through the STNO, the electrons become spin-polarized in the direction of magnetization inside the PL, tunnel through the insulating spacer and exert a torque $\mathbf{T}_S$ on the FL. This torque is called spin-transfer torque (STT) and is proportional to both the angle between local magnetization within each layer and the STT polarization [16], [18]. Consequently, a spin-polarized DC current can be used to cancel out the magnetization damping $\mathbf{T}_\alpha$ through $\mathbf{T}_S$ and excite steady state magnetization precession as it is described in Eq. (1) and Fig.1.(b).

The tunneling magnetoresistance (TMR) effect [19], [20] translates the STT induced magnetization precession into an output voltage with a frequency proportional to the field and DC current, described as [21]:

$$V_{STNO} = (R_{dc} + R_{prec}\cos(2\pi f_{STNO}t + \phi))I_{DC} \quad (2)$$

where R_{dc} is the MTJ DC resistance, R_{prec} is peak precessional resistance induced by the FL magnetization oscillations, f_{STNO} is the operating frequency and ϕ the phase noise. Both R_{prec} and f_{STNO} are strongly affected by changes in both the DC current and externally applied magnetic field H_{ext} (direction and magnitude). Therefore, the external magnetic field can be measured by monitoring the changes in the STNO operating frequency, which have shown a linear relationship [22]. These characteristics offer the possibility of utilizing STNOs as frequency based nanoscaled magnetic field sensors.

B. VCO ADC

Time-based circuits, which process analog signals in time domain rather than voltage domain, have gained substantial interest in recent years. The main driving force has been technology scaling, where signal processing in the voltage domain becomes increasingly challenging at lower supply voltages. In literature, VCO-based ADCs are the most common time-domain circuits studied and developed for both low- and high-frequency applications [13], [14]. Additionally, in recent years time-based amplifiers [23] and neural network circuits [24] have also been proposed.

The conventional single-phase open-loop VCO ADC block diagram is shown in Fig. 2 [25]. A sensor connected to the ADC input generates a voltage $v_{in}(t)$ which is used as a control signal for the VCO. The VCO converts the voltage

Fig. 2. Block diagram of an open-loop VCO-based ADC.

$v_{in}(t)$ to time/phase domain where the conversion relationship (to a first-order approximation) is given by [26]:

$$f_{vco}(t) = f_c + K_{vco}v_{in}(t) \quad (3)$$

where K_{VCO} is the VCO frequency gain and f_c is the free running frequency, defined as the VCO operating frequency when $v_{in}(t) = 0$ V. A digital counter is used to count the number of rising edges within one sampling period at the VCO output, limiting the phase quantization step to 2π for the single-phase architecture shown in Fig. 2. The counter is followed by a differentiator and digital logic, which translate the number of rising edges into a digital code corresponding to $v_{in}(t)$. A quantization error is introduced by the counter since its output is sampled asynchronously relative to the VCO output. However, this quantization error (residual phase) inherently becomes the initial phase of the next sampling period, thus the VCO acts like an integrator. The combination of this time-domain integration, and the delta operation performed by the differentiator resemble a first-order Delta-Sigma ($\Delta\Sigma$) modulator with a noise transfer function (NTF) [25], [26]:

$$NTF(z) = 1 - z^{-1} \quad (4)$$

and a signal transfer function (STF):

$$STF(z) = K_{vco}2\pi \quad (5)$$

in z-domain.

Additionally, the VCO's phase noise has the same NTF as the quantization noise [25], [26]. This is an important feature for the STNO-based magnetic field-to-digital converter since typical STNOs have shown phase noise in the range from -55 dBc/Hz to -72 dBc/Hz at 1 MHz offset, limiting their applications [15]. However, the STNO in the proposed architecture can replace both the sensor and VCO in Fig. 2 and acts as a magnetic field controlled oscillator. This unique characteristic in combination with the inherent noise shaping of open-loop VCO-based ADCs inspired us to explore the proposed architecture.

C. Implementation and simulation setup

The proposed architecture, presented in Fig. 3 consists of a bias-T, low-noise amplifier (LNA), a limiting amplifier chain and a digital core. The challenging aspect of this architecture is the relatively low output power of the emerging STNO devices, requiring amplification to drive the digital counter. Additionally, MTJ STNO measurements have typically shown a tuning range over a few MHz/Oe while macrospin simulations have gone as high as 17 MHz/Oe for an optimized geometry [6]. This wide tunability of the STNO operating frequency requires that the LNA and amplifier chain in Fig. 3 have a wide bandwidth. To meet these challenging requirements, a balun

Fig. 3. The proposed STNO-based magnetic field-to-digital converter.

Fig. 4. Block diagram of the digital core and schematic of the asynchronous counter with double sampling and decoder.

LNA proposed in [27] and shown in Fig. 3 can be used. The balun LNA consists of a common-source (CS) stage and a cross coupled common-gate (CG) stage. This architecture provides an unbalanced to balanced conversion, gain boosting, bandwidth enhancement and good trade-offs in terms of gain balancing, noise performance and matching. The balun LNA is followed by a limiting amplifier chain consisting of five cascode stages to further enhance the gain, and an output CS stage allowing for a large output swing able to drive the digital core [27]. The digital core is presented in Fig. 4 where the first stage consists of a ripple counter and a decoder inspired by the architecture proposed in [14]. Since the ripple counter operates asynchronously relative to the sampling clock, f_s , this architecture is specifically designed to prevent metastability when sampling the counter value. This is achieved by double-sampling the outputs of the frequency divider and employing a series of MUXes to choose sampled values that are guaranteed to be stable. The chosen sampled values are then decoded using xor gates before being passed to the first-order differentiator, finally producing a digital representation of the magnetic field, H_{ext} . It is worth mentioning that frequency dividers in

Fig. 4 are implemented with fully differential latches but are here depicted as single ended for simplicity.

Individual blocks in Fig. 3 have been tested and validated separately by using co-simulations. Specifically, the digital core has been designed at the schematic level and tested together with the behavioral models of other blocks. However, attributed to the extensive simulation times, the system level performance (presented in Section III) is evaluated using Verilog-A/Verilog simulations of analog/digital blocks, without affecting the conclusion in terms of feasibility. The used MTJ STNO model is based on macrospin simulations [28]. The model is based on the MTJ STNO in [29] with a composition: $PtMn(15)/Co_{70}Fe_{30}(2.5)/Ru(0.85)/Co_{40}Fe_{40}B_{20}(2.4)/MgO(0.8)/Co_{60}Fe_{20}B_{20}(1.8)$ (thickness in nm), which provides comprehensive information required for modeling. Additionally, this device has been selected due to its relatively high output power and reasonable phase noise. The STNO macrospin model takes into account critical characteristics, such as phase noise, and frequency and amplitude variation with H_{ext} and I_{DC} [30], [21]. The MTJ STNO model parameters are given in Table I. A DC

TABLE I
STNO AND LNA+AMPLIFICATION CHAIN SIMULATION PARAMETERS

Parameter	Definition	Value
H_{int}	Coupling field	125 Oe
H_A	Anisotropy field	120 Oe
H_s	Demagnetization field	9500 Oe
ϵ (Unitless)	Spin-polarization efficiency	0.7
$q1$ (Unitless)	First coefficient in $\alpha(\xi)$ expansion	20
η (Unitless)	Noise power	0.05
R_{ap}	Antiparallel resistance	650 Ω
R_p	Parallel resistance	300 Ω
ϕ_{ext}	In-plane external field angle	40°
θ_{ext}	Out-of-plane external field angle	0°
α (Unitless)	Gilbert damping parameter	0.018
Δ_t	Time-step for phase noise generation	10 ps
I_{DC}	Applied DC current	1.5 mA
$W \times L$	STNO area	140nm $\times$ 85nm
A_v	LNA+Amplifier chain gain	≈ 35 dB
BW	LNA+Amplifier chain bandwidth	3 – 8 GHz
NF	Noise figure	5 dB
IIP3	Third order input intercept point	-28 dBm

current $I_{DC} = 1.5$ mA was chosen to maximize the output power while not exceeding the STNO breakdown voltage. The measurement results in [29] show a reasonable output power for in-plane angles in the range of $40^\circ < \phi_{ext} < 80^\circ$ (enough to drive the digital core after amplification), while outside this region the power of the STNO fundamental rapidly decreases and the linewidth increases. Consequently, the proposed architecture will not function properly for in-plane field angles outside this range. Therefore, the in-plane and out-of-plane external field angles were chosen as $\phi_{ext} = 40^\circ$ and $\theta_{ext} = 0^\circ$, respectively to achieve a high output power and low phase noise while the rest of the parameters were either extracted from [29] or used as fitting parameters. The MTJ STNO macrospin model does not take into account instantaneous changes in phase noise as a function of H_{ext} [21]. Therefore, in the following simulations the phase noise is assumed to be independent of H_{ext} with a linewidth $\Delta f = 0.15$ GHz. For a more detailed description of the STNO model and parameters in Table I, see [28], [30], [21]. For accurate modeling of the amplifying stages, important characteristics such as gain, bandwidth and noise figure are taken into account. The amplifier simulation parameters, presented in Table I, are inspired by the measurement results of the STNO amplifier in [27]. It is important to note that the bandwidth is between 3 GHz and 8 GHz, which limits the STNO operating frequency to the same range. Lastly, as it was previously mentioned, Verilog descriptions were employed for the digital circuits. However, the functionality of the digital core was also verified up to 8 GHz in a 65nm CMOS technology.

III. RESULTS AND DISCUSSION

The simulated STNO operating frequency, f_{STNO} as a function of the external magnetic field is presented in Fig. 5. The device exhibits a frequency shift of 5 GHz over a change in external field of 700 Oe, resulting in a frequency tunability of $K_{STNO} \approx 7$ MHz/Oe. The relationship is nearly linear for $H_{ext} > 500$ Oe, but for lower external fields it deviates significantly. This non-linearity significantly degrades the performance of the magnetic field-to-digital converter and

Fig. 5. STNO operating frequency as a function of external magnetic field.

Fig. 6. Output spectrum of the VCO-based ADC before and after calibration for $f_{in} \approx 1$ MHz, $f_s = 400$ MHz and NFFT = 4096.

needs to be accounted for. To overcome this issue, a look-up-table (LUT) calibration method was introduced after the differentiator. This method is commonly used to account for non-linearities in open-loop VCO-based ADCs [14], [25]. The effect of the LUT calibration on the output spectrum of the STNO-based magnetic field-to-digital converter is shown in Fig. 6. Attributed to the extremely long simulation time of the STNO macrospin model, a low number of FFT points has been used. The performance of the proposed architecture is evaluated in terms of signal-to-noise and distortion ratio (SNDR), which directly translates to the effective number of bits (ENOB), similar to conventional ADCs. The SNDR as a function of the input bandwidth f_{bw} for different oversampling ratios (OSR), $OSR = f_s/2f_{bw}$, where f_s is the sampling frequency, is presented in Fig. 7. Here the external magnetic field, H_{ext} is assumed to exhibit a sinusoidal shape with a peak value of $H_{ext-p} = 350$ Oe around a DC component of 650 Oe with a frequency of $f_{in} \approx 1$ MHz. The SNDR is relatively independent of OSR for low f_{bw} , suggesting that phase noise dominates the achievable SNDR rather than quantization noise. To verify this assumption, we utilize the analytical equations which describe the VCO-based ADCs [25], [26], where the SNDR takes into account both quantization noise and STNO's phase noise:

$$SNDR = 10 \log_{10} \left(\frac{P_s}{P_{dist} + P_{n,quant} + P_{n,STNO}} \right) \quad (6)$$

where P_{dist} is the distortion, P_s is the signal power defined as:

$$P_s = 0.5 \left(\frac{H_{ext-p} K_{STNO} 2\pi}{f_s} \right)^2 \quad (7)$$

Fig. 7. Simulated and theoretical SNDR as a function of f_{bw} for different f_s , $f_{in} \approx 1$ MHz and $H_{ext-p} = 350$ Oe.

$P_{n,quant}$ is the quantization noise defined as:

$$P_{n,quant} = \frac{\pi^4}{9} \left(\frac{2f_{bw}}{f_s} \right)^3 \quad (8)$$

and $P_{n,STNO}$ is the STNO's phase noise:

$$P_{n,STNO} = \frac{32\pi^2 L \cdot f_{off}^2 f_{bw}}{f_s^2} \quad (9)$$

where L is the phase noise at a frequency offset f_{off} .

In Eq. (9), $L = -53$ dBc/Hz and $f_{off} = 1$ MHz have been chosen for fitting the theoretical SNDR to the simulated SNDR. After introducing the LUT calibration technique, $P_{dist} \approx 0$ in Eq. (6), as it can be seen in Fig. 6 (the spurs are suppressed into the noise floor). The fitting results presented in Fig. 7 demonstrate that the theoretical SNDR match well the simulated SNDR. Additionally, the theoretical expressions allowed us to observe that for low f_{bw} , $P_{n,STNO} \gg P_{n,quant}$, which confirms that the achievable SNDR is limited by phase noise. Consequently, increasing the OSR has negligible effect in the low f_{bw} region. However, as the bandwidth increases, $P_{n,quant}$ also increases as it can be seen from Eq. (8), making SNDR dependent also on quantization noise and OSR. The SNDR as a function of the input power for $f_s = 200$ MHz is presented in Fig. 8 showing a dynamic range of $DR \approx 37$ dB. This dynamic range is limited by the bandwidth of the LNA and amplifier chain (3-8 GHz) as it was previously mentioned. Finally, the simulated SNDR as a function of f_{in} for a full-scale input showed a relatively constant SNDR with f_{in} in the range $0.5 - 5$ MHz, consistent with the theory in Eq. (6)-(9). As it can be seen in Fig. 8, the maximum achievable SNDR is ≈ 38 dB, which translates to:

$$ENOB = \frac{SNDR - 1.76}{6.02} \approx 6 \quad (10)$$

for the proposed architecture. These simulation results validate the feasibility of the proposed magnetic field-to-digital converter and indicate that its performance is limited by the STNO phase noise characteristics.

The performance can be enhanced considering the current development in the STNO technology. For instance, the magnetic field's dynamic range could be tailored to a specific application utilizing other materials and/or by changing the STNO's size. Alternatively, lower phase noise and higher output power

Fig. 8. Simulated SNDR and theoretical SNDR as a function of input power for $f_{in} \approx 1$ MHz, $f_s = 200$ MHz and $f_{bw} = 10$ MHz.

MTJ STNO's have been demonstrated by utilizing vortex oscillations in STNOs [15]. Moreover, vortex based STNOs usually exhibit lower operating frequencies (less than 2 GHz) [15], relaxing the design constraints of the amplifier chain and digital logic. However, they currently show a lower tunability compared to other spintronic oscillators. In recent years, arrays of synchronized spintronic oscillators, which have shown a reduction in phase noise and improvements in output power, have also gained attention [31].

IV. CONCLUSION

In this work, a magnetic field-to-digital converter based on STNOs was proposed. Schematic simulations of individual blocks and behavioral simulations of the whole architecture were carried out, demonstrating its feasibility. The achievable SNDR is fundamentally limited by the STNO phase noise, similar to conventional VCO-based ADCs. However, with the extensive research focused on increasing the output power and reducing the phase noise of spintronic oscillators, the magnetic field-to-digital converter performance could be significantly improved in future implementations.

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