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Postprint

This is the accepted version of a paper published in *IEEE transactions on magnetics*. This paper has been peer-reviewed but does not include the final publisher proof-corrections or journal pagination.

Citation for the original published paper (version of record):

Albertsson, D I. (2019)

Compact Macrospin-Based Model of Three-Terminal Spin-Hall Nano Oscillators

IEEE transactions on magnetics, 55(10): 4003808

<https://doi.org/10.1109/TMAG.2019.2925781>

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Permanent link to this version:

<http://urn.kb.se/resolve?urn=urn:nbn:se:kth:diva-259715>

Compact Macrospin-based Model of Three-terminal Spin-Hall Nano Oscillators

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Emerging spin-torque nano oscillators (STNOs) and spin-Hall nano oscillators (SHNOs) are potential candidates for microwave applications. Recent advances in three-terminal magnetic tunnel junction (MTJ) based SHNOs opened the possibility to develop more reliable and well controlled oscillators thanks to individual spin Hall driven precession excitation and read-out paths. To develop hybrid systems by integrating three-terminal SHNOs and CMOS circuits, an electrical model able to capture the analog characteristics of three-terminal SHNOs is needed. This model needs to be compatible with current electric design automation (EDA) tools. This work presents a comprehensive macrospin-based model of three-terminal SHNOs able to describe the DC operating point, frequency modulation, phase noise and output power. Moreover, the effect of voltage controlled magnetic anisotropy (VCMA) is included. The model shows good agreement with experimental measurements and could be used in developing hybrid three-terminal SHNO/CMOS systems.

Index Terms—Compact model, magnetic tunnel junction (MTJ), spin-Hall nano oscillator (SHNO).

I. INTRODUCTION

SPIN torque nano oscillators (STNOs) are commonly realized with magnetic tunnel junctions (MTJs) composed of an insulator sandwiched between two ferromagnetic layers. By applying a DC current to an MTJ, voltage oscillations in the microwave range can be excited [1]. The DC current gets spin polarized as it passes through the pinned ferromagnetic layer (PL) and transfers its spin angular momentum to the free layer (FL) by exerting an anti-damping torque called spin-transfer torque (STT). STT can compensate the FL intrinsic damping torque, hence inducing a sustainable precession of FL magnetization around the effective magnetic field while the PL has a fixed magnetization direction. Time dependent evolution of the FL magnetization can be described by Landau-Lifshitz-Gilbert equation with an additional Slonczewski spin-transfer term (LLGS) [2], [3]:

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

where $\mathbf{M}$ is the magnetization vector of the FL, γ is the gyromagnetic ratio, $\mathbf{H}_{\text{eff}}$ is the effective magnetic field, $\mathbf{T}_\alpha$ is the intrinsic damping term and $\mathbf{T}_S$ is the anti-damping term describing STT induced by the spin current. Because of tunneling magnetoresistance (TMR) [4], [5], the magnetization oscillations described by (1) change the MTJ resistance, leading to voltage oscillations as output. Moreover, the frequency of oscillations and the output power of STNOs can be tuned by changing either the DC current or the effective magnetic field. The wide frequency tunability [6], [7] and integration compatibility with CMOS circuits make STNOs an interesting nano-device for a number of applications, e.g. as microwave oscillators in communication systems [8], [9], neuromorphic computing [10], wireless current sensors [11] and magnetic

Fig. 1. Three-terminal MTJ-SHNO architecture.

field detection [12]. However, the major drawback of STNOs is currently their low output power and substantial phase noise [1]. A possible solution to overcome these drawbacks is utilizing mutually synchronized STNOs or spin-Hall nano oscillators (SHNOs) where an increase in the output power and a reduction in phase noise have been observed [13]–[15]. An MTJ based three-terminal SHNO (MTJ-SHNO) as shown in Fig. 1 [16], is composed of an MTJ like STNOs, however an additional heavy-metal (HM) strip is placed adjacent to the FL. Rather than inducing and detecting oscillations with current through the MTJ as in STNOs, these tasks have been separated to two independent ports. While the current through the HM strip induces auto-oscillations, a small current is applied to the MTJ for readout. This allows for independent modulation of operating frequency and detection of resistance oscillations

[16]. Moreover, with the three-terminal architecture, the MTJ dielectric breakdown is avoided due to the relatively small MTJ readout current density [17], [18]. Considering these advantages, three-terminal MTJ-SHNOs could be a promising candidate for next generation hybrid spintronic-CMOS architectures, e.g. in neuromorphic or associative computing [19], [20]. To enable such developments, a comprehensive model able to capture the analog characteristics of three-terminal MTJ-SHNOs is needed. The model needs to capture fundamental characteristics including the DC operating point, frequency modulation/generation, phase noise and output power of three-terminal MTJ-SHNOs. Moreover, the model needs to be compatible with current CMOS electric design automation (EDA) tools and perform simulations in a reasonable amount of time. Similar models have been proposed for STNOs [21]–[24] but a comprehensive model of three-terminal MTJ-SHNOs able to capture the previously mentioned characteristics is missing.

This paper presents a compact macrospin-based model of three-terminal MTJ-SHNOs. The model is implemented in Verilog-A and validated against experimental measurements performed in [16]. The model development is based on the device in [16], where both spin current and voltage-controlled magnetic anisotropy (VCMA) are utilized for frequency modulation. Specifically, the model is developed for in-plane magnetized three-terminal MTJ-SHNOs.

The following sections are organized as follows. Section II describes the theory and model development. Section III presents the simulation results and comparison to experimental data, while Section IV concludes the paper.

II. COMPACT THREE-TERMINAL MTJ-SHNO MODEL

The proposed three-terminal MTJ-SHNO model is based on a previously developed STNO model in [21], [22]. Both models are largely based on the nonlinear auto-oscillator theory of microwave generation by spin-polarized current proposed by Slavin and Tiberkevich [3], which is an analytical approach to describe STT induced dynamics. Their approach allows for relatively fast simulations times while achieving a reasonable accuracy. The specific contribution of the three-terminal model compared to the previous STNO model in [21], [22] are: 1) The model is adapted for three-terminal MTJ-SHNOs, 2) the effect of PMA and VCMA is included and, 3) both the fundamental and second harmonic of oscillations are considered.

A. Basic operation of three-terminal MTJ-SHNOs with VCMA

When a charge current is passed through a HM with high spin-orbit interaction, notable transverse spin-current is accumulated at HM surfaces [25]. This phenomena is called spin-Hall effect (SHE) [26] and is crucial for implementation of MTJ-SHNO architectures. In certain materials (e.g. Pt [13], Ta [27] and W [28], [29]), the conversion ratio from charge to spin current, called spin-Hall angle (SHA) is large enough to exert a sufficient torque on an adjacent FL to excite magnetization precession. The three-terminal MTJ-SHNO architecture

proposed in [16] is based on these principles where a Ta strip is providing spin injection into the FL. Although the purpose of the MTJ current is primarily to perform read-out, it can also be used to tune perpendicular magnetic anisotropy (PMA) [30], [31]. The advantages of utilizing PMA and its tunability include: 1) The PMA partially cancels the demagnetization field, leading to lower operating current and 2) tunability introduces an additional degree of freedom as it directly modifies the operating frequency. To realize these characteristics, the MTJ in [16] is designed to have a large interfacial anisotropy energy between the HM/ferromagnetic/insulator, leading to a PMA. The PMA is then modulated through the electric field induced by the MTJ current. This change in PMA, called voltage controlled magnetic anisotropy (VCMA) can be explained in terms of the Rashba spin-orbit coupling and change in occupancy of atomic orbitals in thin (<2 nm) ferromagnetic layers [31]–[33]. Thus, when developing the three-terminal MTJ-SHNO model, these phenomena need to be included for accurate modeling.

B. Effective Magnetic Field

The first step in obtaining the characteristics of a three-terminal MTJ-SHNO device is to calculate the effective magnetic field H_{eff} inside the FL. In the presence of an externally applied magnetic field H_{ext} with the in-plane angle ϕ_{ext} relative to the $\hat{x}$ axis and out-of-plane angle θ_{ext} relative to the $\hat{z}$ axis, the following set of equations can be derived in $\hat{x}$, $\hat{y}$ and $\hat{z}$ direction [3], [21]:

$$H_{eff} \cos(\theta_{eff}) \cos(\phi_{eff}) = H_{ext} \cos(\theta_{ext}) \cos(\phi_{ext}) + H_A \cos(\theta_{eff}) \cos(\phi_{eff}) - H_{int} \quad (2a)$$

$$H_{eff} \cos(\theta_{eff}) \sin(\phi_{eff}) = H_{ext} \cos(\theta_{ext}) \sin(\phi_{ext}) \quad (2b)$$

$$H_{eff} \sin(\theta_{eff}) = H_{ext} \sin(\theta_{ext}) - H_{eff}^{demag} \sin(\theta_{eff}) \quad (2c)$$

where H_{eff} is the effective magnetic field, θ_{eff} and ϕ_{eff} are in-plane and out-of-plane effective angles, H_A is the anisotropy field, H_{int} is the inter-layer coupling field and H_{eff}^{demag} the effective demagnetization field.

The effective demagnetization field consists of contributions from the demagnetization field and the PMA. As it has been mentioned, the magnitude of the PMA depends on the electric field induced by the voltage drop across the MTJ, allowing for the VCMA. Taking this into account, the effective demagnetization field can be described as [16]:

$$H_{eff}^{demag} = H^{demag} - H_{PMA} + \frac{dH_{PMA}}{dV_{MTJ}} V_{MTJ} \quad (3)$$

where $H^{demag} = 4\pi M_s$ is the demagnetization field, M_s is the saturation magnetization, H_{PMA} is PMA, dH_{PMA}/dV_{MTJ} is the rate of PMA change in Oe/V (the VCMA) and V_{MTJ} is the DC voltage drop across the MTJ. For the three-terminal device proposed in [16], the externally applied magnetic field H_{ext} is applied in-plane ($\theta_{ext} = 0$), this simplifies (2a)–(2c) to [21]:

$$H_{ext} \sin(\phi_{ext}) \frac{\cos(\phi_{eff})}{\sin(\phi_{eff})} + H_{int} = H_A \cos(\phi_{eff}) + H_{ext} \cos(\phi_{ext}) \quad (4a)$$

$$H_{eff} = H_{ext} \frac{\sin(\phi_a)}{\sin(\phi_{eff})} \quad (4b)$$

where H_{eff} and ϕ_{eff} can easily be found. As can be observed, H_{eff}^{demag} and the VCMA no longer contribute to H_{eff} . However, H_{eff}^{demag} affects the operating frequency as will be shown in the following sub-section.

C. DC characteristics

Fig. 2. Equivalent DC model of the three-terminal MTJ-SHNO.

The DC operating point is a key parameter for accurate modeling of three-terminal MTJ-SHNOs and simulations together with CMOS circuits. The DC model is presented in Fig. 2, where the MTJ is assumed to be located in the center of a HM strip. Consequently, the DC operating point can be defined as:

$$V_{DC} = V_{MTJ} + \frac{R_{HM}}{2}(I_{MTJ} + I_{HM}) \quad (5)$$

where $V_{MTJ} = I_{MTJ}R_{MTJ-DC}$, R_{HM} is the HM strip resistance and I_{HM} is the current through the HM. All the parameters in (5) excluding the MTJ DC resistance R_{MTJ-DC} are known constants. R_{MTJ-DC} depends on the minimum resistance of the device R_p (when the magnetization of the free and fixed layers are parallel) and maximum (anti-parallel) R_{ap} [4] [5]. This resistance change between parallel and anti-parallel magnetization configuration is quantified in terms of the tunneling magnetoresistance (TMR) ratio defined as:

$$TMR = \frac{R_{ap} - R_p}{R_p} \quad (6)$$

The TMR has shown a strong dependence on the voltage across the MTJ, V_{MTJ} . Moreover, this dependence is asymmetric with respect to the sign of V_{MTJ} [34]. This behavior can be described with the fitting function [35]:

$$TMR(V_{MTJ}) = \begin{cases} TMR_0 \frac{1}{1+(V_{MTJ}/V_h^+)^2}, & V_{MTJ} \geq 0 \\ TMR_0 \frac{1}{1+(V_{MTJ}/V_h^-)^2}, & V_{MTJ} < 0 \end{cases} \quad (7)$$

where $TMR(V_{MTJ})$ is the TMR at a voltage V_{MTJ} applied across the MTJ, TMR_0 is the TMR at $V_{MTJ} = 0$ and V_h^+ , V_h^- are fitting parameters for positive and negative voltages. The fitting parameters correspond to the value of V_{MTJ} when $TMR(V_{MTJ}) = 0.5TMR_0$. To accurately model three-terminal MTJ-SHNOs, equation (7) needs to be modified. The scaling of TMR described by (7) is directly related to a decrease in both R_{ap} and R_p as a function of V_{MTJ} . Since the device is biased by the current I_{MTJ} which induces the

voltage V_{MTJ} , it is convenient to fit R_p and R_{ap} as a function of I_{MTJ} . Moreover, it is critical to model how R_p and R_{ap} scale for both positive and negative current values. For the anti-parallel resistance R_{ap} this leads to:

$$R_{aps}(I_{MTJ}) = \begin{cases} \frac{R_{ap}}{1+(I_{MTJ}/I_{ap-h}^+)^{N_f}}, & I_{MTJ} > 0 \\ \frac{R_{ap}}{1+(I_{MTJ}/I_{ap-h}^-)^{N_f}}, & I_{MTJ} < 0 \end{cases} \quad (8)$$

where R_{aps} is the scaled anti-parallel resistance, R_{ap} is anti-parallel resistance when $I_{MTJ} = 0$, I_{ap-h}^+ (I_{ap-h}^-) is a fitting parameter for positive (negative) MTJ current and N_f is a fitting parameter (between 1-2) defining the steepness of the fitted curve. The decrease in parallel resistance can be written in a similar way:

$$R_{ps}(I_{MTJ}) = \begin{cases} \frac{R_p}{1+(I_{MTJ}/I_{p-h}^+)^{N_f}}, & I_{MTJ} > 0 \\ \frac{R_p}{1+(I_{MTJ}/I_{p-h}^-)^{N_f}}, & I_{MTJ} < 0 \end{cases} \quad (9)$$

where R_{ps} is the scaled parallel resistance, R_p is parallel resistance when $I_{MTJ} = 0$ and I_{p-h}^+ (I_{p-h}^-) is a fitting parameter for positive (negative) MTJ current. Although 5 fitting parameters are needed to describe this behaviour, they are critical for accurate modeling of three-terminal MTJ-SHNOs. Moreover, the DC resistance, R_{MTJ-DC} , has experimentally shown a squared-sine dependence on ϕ_{eff} as [36]:

$$R_{MTJ-DC}(\phi_{eff}) \approx R_{ps} + (R_{aps} - R_{ps})\sin^2\left(\frac{\phi_{eff}}{2}\right) \quad (10)$$

As it can be observed from (8), (9) and (10), the DC operating point can vary significantly as a function of both I_{MTJ} and ϕ_{eff} .

D. Operating Frequency

To capture the frequency behavior of three-terminal MTJ-SHNOs, the theory developed in [3] is utilized. Within this theory, spin based oscillators are described with the universal model of an auto-oscillator with nonlinear frequency shift and negative damping. The frequency of operation ω_p can then be defined as [3]:

$$\omega_p \approx \omega_0 + N\bar{p} \quad (11)$$

where ω_0 is the ferromagnetic resonance (FMR) frequency, N is the nonlinear frequency shift coefficient and $\bar{p}$ is the average dimensionless spin-wave power. To compute ω_p , these parameters need to be obtained according to the theory in [3]. The FMR frequency ω_0 can be calculated according to the Kittel equation (assuming $\theta_{ext} = 0$) [3]:

$$\omega_0 = \gamma\sqrt{(H_{eff} - \sin^2\phi_{eff}H_A)(H_{eff} + H_{eff}^{demag})} \quad (12)$$

The dimensionless spin-wave power $\bar{p}$ is a critical parameter that partially describes both the damping and STT induced anti-damping in the FL (T_α and T_S in (1)). According to [3, eq. (84b)], $\bar{p}$ depends on the nonlinear damping coefficient Q , the effective noise power η and the supercriticality parameter ζ . As it is detailed in [3], Q is defined in terms of q_1 , a parameter that can be obtained from experimental measurements. This is a cumbersome task in reality and Q will be considered as

a phenomenological parameter. Similarly, the effective noise-power η is a fitting parameter that together with Q has to be tuned to fit the experimental data [21]. The supercritical parameter ζ can be defined as [3]:

$$\zeta = \frac{I_{HM-eff}}{I_{th}} \quad (13)$$

where I_{th} is the threshold current needed to excite oscillations in the device and I_{HM-eff} is the effective current in the HM strip. The MTJ current density is assumed low, thus STT induced by I_{MTJ} as it passes through the MTJ is considered insignificant. However, as I_{MTJ} passes into the HM, it generates an additional spin-current on the HM surface due to the SHE. Consequently, there is asymmetry in the spin-current density on each side of the MTJ [37]. Since this additional current affects half of the MTJ area, we define I_{HM-eff} as:

$$I_{HM-eff} = I_{HM} + \frac{I_{MTJ}}{2} \quad (14)$$

where an assumption is made that additional spin-current contribution from I_{MTJ} can be modeled as an evenly distributed current in the HM with half the magnitude. In (13) the threshold current I_{th} is defined in terms of the damping term Γ_G and the coefficient σ as:

$$I_{th} = \frac{\Gamma_G}{\sigma} \quad (15)$$

where Γ_G can be obtained with [3, eq. (104b)]. The coefficient σ for STNOs is defined in [3] in terms of the spin-polarization efficiency ϵ . For three-terminal MTJ-SHNOs, this coefficient needs to be redefined in terms of the effective SHA θ_{SH} as [38]:

$$\sigma = \frac{\theta_{SH} g \mu_B}{2e A_{HM} t_f M_{eff}^{demag}} \cos(\phi_{eff}) \quad (16)$$

where g is the spectroscopic Lande factor, μ_B is the Bohr magneton, e is the electron charge, A_{HM} is the HM cross sectional area, t_f is the FL thickness and M_{eff}^{demag} is the effective demagnetization.

Lastly, to calculate the operating frequency with (11), the nonlinear frequency shift coefficient, N , can be directly computed with [3, eq. (105a)] and coefficients in [3, eqs. (106a)-(106h)].

E. Linewidth

Thermal fluctuations generates phase noise in the output signal of STNOs and SHNOs. The phase noise is quantified in terms of the linewidth, defined as the full frequency spectrum at half of the maximum power. These fluctuations are one of the major drawbacks of STT based oscillators and need to be accounted for in a realistic model. In [3], the behavior of phase noise is divided into three regions: sub-threshold ($I_{HM-eff} < I_{th}$), near threshold ($I_{HM-eff} \approx I_{th}$) and above threshold ($I_{HM-eff} > I_{th}$).

In the below threshold regime, $\bar{p}$ approaches the effective noise power η , and the phase noise behaves linearly. The full linewidth $2\Delta\omega$ can then be approximated as [3]:

$$2\Delta\omega = 2\Gamma_G(1 - \zeta) \quad (17)$$

where it can be observed that the linewidth decreases linearly as a function of I_{HM-eff} . In the above threshold regime, the linewidth can be analytically approximated in two limiting cases. First, when both the temperature T and the linewidth are sufficiently small, the power spectrum has a Lorentzian shape. The linewidth can then be approximated as [3]:

$$2\Delta\omega = (1 + \nu^2)\Gamma_+(p_0) \frac{k_B T}{\epsilon(p_0)} \quad (18)$$

where ν is the normalized dimensionless nonlinear frequency shift obtained with [3, eq. (33)], $\Gamma_+(p_0) = \Gamma_G(1 + Qp_0)$, where p_0 is the stationary power calculated with [3, eq. (24)], k_B is the Boltzman constant, T is temperature and $\epsilon(p_0)$ is the stationary oscillator energy calculated with [3, eq. (77)]. Alternatively, the other limiting case is when the linewidth is relatively large. In this case the linewidth is approximated as [3]:

$$2\Delta\omega = 2|\nu| \sqrt{\Gamma_+(p_0)\Gamma_p} \sqrt{\frac{k_B T}{\epsilon(p_0)}} \quad (19)$$

where Γ_p is the effective damping rate for small power deviations obtained with [3, eq. (27b)]. For STNOs in [3], [21], the calculated linewidth using (18) shows a better resemblance to experimental data compared to (19), thus (18) will be employed in the above-threshold region. Lastly, in the near threshold region, the linewidth in (17) approaches zero (when $\zeta \approx 1$) which is not a realistic behavior. Moreover, as is discussed in [3], it is difficult to obtain an analytical expression for the linewidth in this region. Thus, the same approach as in [21] is applied and it is assumed that in the near threshold region, corresponding to $0.85I_{th} < I_{HM-eff} < I_{th}$, (18) can describe $2\Delta\omega$ to a reasonable degree.

F. Output Power

To approximate the output power of three-terminal MTJ-SHNOs, equations that have been experimentally validated for STNOs in [6] and [39] are employed. In [21] the same equations were used for STNOs and showed good match with experimental results. Moreover, in the three-terminal MTJ-SHNO model, the power of both the fundamental and second harmonic are included. This allows for simulations that capture the intrinsic frequency doubling utilized in [16] to get a higher operating frequency [36]. By rewriting the equations describing the fundamental and second harmonic peak power in [6], the peak oscillation resistance R_{p1} of the fundamental and R_{p2} of the second harmonic can be defined as [21]:

$$R_{p1}(\omega_p) = \left(\frac{R_{aps} - R_{ps}}{R_0} \right)^2 J_1^2(\phi_{prec}) \sin^2(\phi_{eff}) \frac{R_0}{8} \quad (20a)$$

$$R_{p2}(2\omega_p) = \left(\frac{R_{aps} - R_{ps}}{R_0} \right)^2 J_2^2(\phi_{prec}) \cos^2(\phi_{eff}) \frac{R_0}{8} \quad (20b)$$

where $\phi_{prec} = 2\sin^{-1}(\sqrt{p})$ is the precision angle [39], R_0 is defined as [6]:

$$R_0 = \left(\frac{R_{aps} + R_{ps}}{2} \right) - \left(\frac{R_{aps} - R_{ps}}{2} \right) J_0(\phi_{prec}) \cos(\phi_{eff}) \quad (21)$$

and $J_0(\phi_{prec})$, $J_1(\phi_{prec})$, $J_2(\phi_{prec})$ are Bessel functions. To implement the Bessel function in Verilog-A, approximations in terms of trigonometric functions are employed [40]. Finally, the model's output signal can be defined by combining (5), (11) and (20):

$$V_{out} = V_{DC} + R_{p1}(\omega_p) \cos(\omega_p t + \psi) I_{MTJ} + R_{p2}(2\omega_p) \cos(2\omega_p t + \psi) I_{MTJ} \quad (22)$$

where ψ is the phase noise. To generate phase noise in the output signal, the approach developed in [22] is utilized. This method is based on the assumption that phase noise is dominated by Gaussian white noise with a zero mean and a standard deviation of $\sqrt{2\Delta\omega/\Delta t}$, where Δt is a virtual simulation step. The phase noise is then updated periodically in intervals of Δt to generate a continuous linear phase change.

III. MODEL VERIFICATION

Throughout this section, the Verilog-A model is compared against experimental measurements performed in [16], where the three-terminal MTJ-SHNO is composed of a $Ta(6)/Co_{40}Fe_{40}B_{20}(1.5)/MgO(1.2)/Co_{40}Fe_{40}B_{20}(4)/Ta(5)/Ru(5)$ stack (thickness presented in nm). The corresponding model parameters are listed in Table I. It is worth mentioning that a relatively large value of $Q = 10$ was chosen (usually, for STNOs $0 < Q < 3$ [3]) due to the weak frequency shift as a function of I_{HM-eff} .

In [16], the three-terminal device is biased in the parallel state with an in-plane angle of $\phi_{ext} = 0^\circ$. By doing this, the fundamental peak should be completely suppressed while the second harmonic reaches a maximum output power (called intrinsic frequency doubling) [36]. Measurement results presented in [16] show a weak fundamental is observed in [16, FIG. 2 (a)], indicating a small misalignment between the PL and FL, according to (20). Consequently, for the following simulations where I_{HM} is swept (while $I_{MTJ} = 60\mu A$), the external field angle is adjusted to $\phi_{ext} = 2^\circ$. In Fig. 3, the simulated frequency shift, linewidth, integrated power and peak power of the second harmonic are presented. For current values below 0.2 mA, the output power reaches the noise floor, indicating $I_{th} \approx 0.2mA$. Thus, values below 0.2 mA were not considered in [16]. The operating frequency in Fig. 3 (a) is accurately captured for current values $I_{HM} > 0.3mA$, but deviates slightly at lower current values. Moreover, as it can be observed in Fig. 3 (c), the integrated power is slightly underestimated. This error might be related to the assumption made in (14) where the magnitude of additional spin-current generated by I_{MTJ} is only considered, not the asymmetry. This asymmetry could have other consequences not captured by a macrospin model. Consequently, this limits the accuracy of a macrospin approach; however it still provides a reasonable fit for developing MTJ-SHNO/CMOS circuits.

TABLE I
MODEL PARAMETERS USED FOR COMPARISON WITH MEASUREMENTS IN [16]. THE COORDINATE SYSTEM IS DEFINED SUCH THAT NEGATIVE VALUES IN [16] ARE PRESENTED AS POSITIVE VALUES.

Parameter	Symbol	Value	Unit
Parallel resist.	R_p	4465	Ω
Anti-parallel resist.	R_{ap}	5150	Ω
HM resistance	R_{HM}	1800	Ω
HM thickness	t_{HM}	6	nm
HM width	w_{HM}	1.2	μm
MTJ width	w_{MTJ}	180	nm
MTJ length	l_{MTJ}	50	nm
FL thickness	t_f	1.5	nm
HM current	I_{HM}	0.8	mA
MTJ current	I_{MTJ}	60	μA
R_{aps} fitting param.	I_{ap-h}^+/I_{ap-h}^-	450/-360 [†]	μA
R_{ps} fitting param.	I_{p-h}^+/I_{p-h}^-	800/-750 [†]	μA
Slope fitting param.	N_f	1.5 [†]	-
In-plane angle	ϕ_{ext}	2/8*	Deg.
External field	H_{ext}	160	Oe
Coupling field	H_{int}	76*	Oe
Anisotropy field	H_A	0	Oe
PMA	H_{PMA}	11700*	Oe
Demag. field	H^{demag}	13000	Oe
VCMA coeff.	dH_{PMA}/dV_{MTJ}	740	Oe/V
Nonlinear damping	Q	10 [†]	-
Effective noise	η	0.08 [†]	-
Gilbert damping	α_G	0.05 [†]	-
SHA	θ_{SH}	0.15	-

Parameter values without notes are measured/suggested in [16]

*Slightly adjusted compared to measured/suggested values in [16].

[†]Fitting parameters.

Next, the effect of varying I_{MTJ} while keeping $I_{HM} = 0.8mA$ is investigated. By comparing [16, FIG. 2(a)] and [16, FIG. 3(a)] a much stronger fundamental peak is observed for the same measurement ($I_{MTJ} = 60\mu A$, $I_{HM} = 0.8mA$) in the latter. This might be related to a misalignment in ϕ_{ext} between the two measurements (according to (20)). Thus, for the following simulations where I_{MTJ} is swept, ϕ_{ext} is adjusted to a value of 8° . Comparison between simulated and experimental data is presented in Fig. 4. It is worth emphasizing that these changes in the operating frequency, linewidth and power are not caused by STT, but are induced by the $VCMA \propto I_{MTJ}$ as is described by (3). The frequency shift as a function of I_{MTJ} can be understood by considering (12) where a decrease in PMA leads to a higher ω_0 and through (16) to a decreased negative frequency shift $N\bar{p}$ (and the opposite for a lower PMA). The combination of these two effects leads to the substantial frequency shift of $\approx 0.4GHz$ shown in Fig. 4 (a). Experimental measurements on the linewidth $2\Delta\omega$ as a function of MTJ current were not performed in [16], however a slight reduction is observed for the simulated linewidth in Fig. 4 (b). This reduction in linewidth is directly related to an increase in the oscillation power $\epsilon(p_0)$ in (18). In Fig. 4 (c), the simulated integrated power closely follows the experimental data. However, the peak power in Fig. 4 (d) is slightly overestimated. This mismatch is most likely related to the simplification made in (14) as it was discussed in regards to Fig. 3 (c). It is worth emphasizing that the peak power can vary slightly between simulations. The method used to

Fig. 3. Three-terminal MTJ-SHNO characteristics as a function of I_{HM} : (a) operating frequency, (b) linewidth, (c) integrated power and (d) peak power of the second harmonic.

generate phase noise in the transient signal is the cause of this, where a random number generator is used to emulate phase noise [22]. This explains the uneven increase in peak power as a function of I_{MTJ} .

Additionally, the effect of sweeping the external field H_{ext} from 125 Oe - 190 Oe, while keeping $I_{MTJ} = 60\mu A$ and $I_{HM} = 0.8mA$ constant, was simulated. The results and comparison with experimental data is presented in Fig. 5. This frequency shift is due to changes in the ferromagnetic resonance ω_0 as a function of the external field H_{ext} , as is described by (12). Lastly, the transient output waveform of the model for an in-plane angle of $\phi_{ext} = 2^\circ$ is presented in Fig. 6. Here the effect of phase noise can be clearly seen as a distortion in the output signal, allowing for realistic transient simulations of the three-terminal MTJ-SHNO with CMOS circuits.

Fig. 4. Three-terminal MTJ-SHNO characteristics as a function of I_{MTJ} : (a) operating frequency, (b) linewidth, (c) integrated power and (d) peak power of the second harmonic.

IV. CONCLUSION

A macrospin based model of the three-terminal MTJ-SHNO which includes phase generation and VCMA has been developed. The model was implemented in Verilog-A and compared to experimental measurements. The simulated values show a relatively good resemblance to the measured data despite the model simplicity. Although there are minor differences between simulations and experimental data, the model allows for design and simulations of hybrid three-terminal MTJ-SHNO/CMOS circuits and could accelerate the widespread use of spintronic devices in next generation electronics.

ACKNOWLEDGMENT

This work was supported by the Swedish Research Council VR.

Fig. 5. Operating frequency of second harmonic as a function of the external field.

Fig. 6. Transient output waveform of the model for $\phi_{ext} = 2^\circ$, the DC offset has been filtered out for clarity.

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