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An Analogue Baseband Chain for a Magnetic Tunnel Junction Based RF Signal Detector

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Abstract—This work presents an analogue baseband (BB) chain for a magnetic tunneling junction (MTJ) based radio-frequency (RF) signal detector fully integrated in a hybrid CMOS-MTJ technology. The BB chain contains a 6th-order g_m -C low-pass filter (LPF), a BB amplifier, a comparator, and a current bank. According to measurement results, the 6th-order LPF with a cut-off frequency of 10 MHz consumes a very low DC power of 2.41 mW. Its DC power consumption per pole of 0.4 mW is the lowest among the state-of-the-art LPFs. The LPF can be also switched on and off very fast within 110 ns. With the fast switchability and the low power consumption, the LPF outperforms the state of the art. Furthermore, the complete BB chain can transform a 2.5 mV_{pp}, 5 Mbps BB signal into digital data with a bit error rate fewer than 10^{-6} . The BB chain consumes 2.85 mW including all bias circuits. To achieve power efficiency, the BB chain is designed to operate under an aggressive duty-cycling mode. The switch-on time of the BB chain is within 200 ns.

Index Terms—Spintronic, magnetic tunnel junction, low-pass filters, CMOS integrated analogue circuits, frequency detector

I. INTRODUCTION

A. Magnetic Tunnel Junction

Magnetic tunnel junctions (MTJ) are nanometre-scale devices that typically consist of multiple ferromagnetic layers separated by a nonmagnetic layer. There are two characteristics recently discovered in the MTJs making them very attractive in RF-related applications. The first one is the DC-to-RF convertibility with which an MTJ can operate naturally as a current controlled oscillator (CCO) [1]. The second one is the RF-to-DC property, hence an inverse phenomenon to the first one, known as the spin-torque diode effect [2]. In this case, when the DC current is suppressed and an RF signal is injected with a frequency that conforms to the intrinsic oscillation frequency of the MTJ, the MTJ outputs a DC signal. Recent studies have shown that the RF signal detection sensitivity (DS) of MTJ-based diodes can be higher ($DS \approx 10^4$ mV/mW, [3]) and much higher ($DS \approx 10^5$ mV/mW, [4]) than CMOS diodes ($DS \approx 4 \cdot 10^3$ mV/mW), thereby paving a lightning path for MTJ-based RF detection in telecommunication markets.

B. MTJ Based RF Signal Detector

It has been demonstrated that MTJs standalone can function as both the CCO and RF detector [5] [6] [7] [8] [9]. A milestone step that needs to be taken is integration of the MTJs into conventional CMOS technologies, which has been carried out within the frame of GREAT project [10]. One of the project goals is to demonstrate a fully integrated MTJ based RF signal detector where the RF-to-DC functionality of the MTJs can be

Fig. 1. Principle diagram of the MTJ based RF signal detector

fully exploited. Fig. 1 shows the block diagram of the newly developed MTJ based RF signal detector.

Here an MTJ array similar to the one introduced in [10] is implemented as the core of the RF detection. Since each MTJ is designed to respond to a single and narrow RF band, the major advantage of the MTJ based receiver compared to a conventional direct conversion receiver lies in the fact there is no need of an RF band-pass filter, thus largely reducing the design complexity, silicon area and cost. Our newly developed MTJ based RF signal detector is targeted to operate in a truly multiple-band standard from 1 to 10 GHz.

The prototype circuit is fabricated in a Towerjazz (TJ) hybrid CMOS-MTJ technology. The active device layer and the first two metal layers, metal layer 1 and 2, are implemented in a TJ 180 nm CMOS technology. The MTJ layers are located above metal layer 2. Above the MTJ layers, metal 3 and 4 layers are fabricated in a TJ 130 nm CMOS technology.

C. Analogue BB Chain

The first step that has been accomplished and is to be presented in this work is the design and implementation of an analogue BB chain for the MTJ-based frequency detector. As shown in Fig. 1, the BB chain, which consists of a LPF, a BB amplifier, a comparator and a current bank, transforms the BB signal generated by the MTJ into a digital bit stream with a maximal data rate below 10 Mbps. Since the output voltage of an MTJ lies in the range of several millivolts to several ten millivolts [4], the BB chain is specified to have a sensitivity of -50 dBm (2 mV_{pp}) at a 50 Ω system. In addition, to achieve power efficiency, the BB chain is designed to work in an aggressive duty-cycling mode with a duty cycle fewer than 0.1%. A fast settling time T_s of the BB chain is specified to lie within 200 ns.

II. CIRCUIT DESIGN

A. Design of a 6th-Order LPF

To eliminate both the input RF signal, noise and simultaneously enhance the weak BB signal generated by the MTJ, the LPF is targeted to have a cut-off frequency f_c of 10 MHz and a transition steepness of -120 dB/decade. To avoid the shift of f_c and pass-band voltage gain A_{LP} due to process variations in fabrication, f_c should be adjustable from 5 MHz to 30 MHz, as well as A_{LP} from 0 to 20 dB. Moreover, the LPF should settle very fast, which is specified within 200 ns. At last, the LPF should also have a low power consumption P_{DC} of < 4 mW.

To realize the filter characteristic of -120 dB/decade, we designed a 6th-order LPF that consists of three cascaded 2nd-order LPFs with the same topology. The transfer function of the 6th-order LPF is thus related to the transfer function of the 2nd-order LPF:

$$G_{LP,6th}(s) = (G_{LP,2nd}(s))^3 \quad (1)$$

1) Tow-Thomas Topology Based OTA-C 2nd-Order LPF:

Our design focus is now shifted to realization of the 2nd-order LPF with a fast T_S and low P_{DC} . As is shown in Fig. 2, here we choose a g_m -C topology named after Tow-Thomas (TT) for our 2nd-order LPF [11] [12]. According to the studies in [12], the TT topology compared to other topologies such as multiple feedback loop provides following benefits: lower transconductance (g_m) required for its OTAs, lower DC power consumption, faster settling and less mismatch due to no resistors required in the signal path. From Fig. 2, we can derive the filter's transfer function $G_{LP,2nd}(s) = \frac{V_{out}}{V_{in}}$:

$$G_{LP,2nd}(s) = \frac{A_{LP,2nd}\omega_0^2}{s^2 + \frac{\omega_0}{Q}s + \omega_0^2} = \frac{\frac{g_{T2}g_{T4}}{C_1C_2}}{s^2 + \frac{g_{T3}}{C_2}s + \frac{g_{T1}g_{T2}}{C_1C_2}} \quad (2)$$

where its cutoff frequency $f_{c,2nd}$, voltage gain $A_{LP,2nd}$, and quality factor Q can be described as the following formula:

$$f_{c,2nd} = \frac{1}{2\pi} \sqrt{\frac{g_{T1}g_{T2}}{C_1C_2}}; A_{LP,2nd} = \frac{g_{T4}}{g_{T1}}; Q = \frac{\sqrt{g_{T1}g_{T2}}}{g_{T3}} \sqrt{\frac{C_2}{C_1}} \quad (3)$$

For the sake of design simplicity, we choose $g_{T1} = g_{T2} = g_{T3} = g_{T0}$. Thus, Eq. (3) can be simplified into:

$$f_{c,2nd} = \frac{1}{2\pi} \frac{g_{T0}}{C_0}; A_{LP,2nd} = \frac{g_{T4}}{g_{T0}}; Q = \sqrt{\frac{C_2}{C_1}} \quad (4)$$

Fig. 2. Schematic of the TT-based 2nd-order LPF

Via Eq. (4) we can first see that the filter's quality factor is independent of its $f_{c,2nd}$ and $A_{LP,2nd}$. Moreover, we design here $Q = \sqrt{2}/2$ due to the compromise between a relatively fast settling time and a small overshoot effect for an under-damped 2nd-order filter. This leads to $C_1 = 2C_2 = 2C_0$. Furthermore, we can tune $f_{c,2nd}$ by only changing g_{T0} . To make the gain $A_{LP,2nd}$ tunable, g_{T4} is designed to have a different digital tuning unit from g_{T0} .

2) OTA Design: As is shown in Fig. 3, the proposed OTA consists of an input g_m stage, an output impedance enhancement stage, a digital tuning stage, a common mode feedback (CMFB) loop and several power-disable switches. The input and output stages together build up a well-known current mirror cascode topology which provides the benefit to tune g_m with ease. g_m can be defined by the DC output current of the output stage which is adjustable via the digital tuning unit. The output current of the positive side of the output stage $I_{o,+}$ consists of the current flowing through M_{3a} and the current flowing through all the PMOS branches of the digital unit, which can be described as:

$$I_{o,+} = I_{D,3a} + \sum_{i=0}^{n-1} b_i 2^{i-1} I_{D,3a} = I_{IN} + \sum_{i=0}^{n-1} b_i 2^{i-1} I_{IN} \quad (5)$$

where b_i refers to the i^{th} digital bit value, either 1 or 0, that determines the on-off state of the PMOS switches $M_{13,14}$. In addition, n is the total number of the digital branches that are connected in parallel. For the OTAs g_{T1} , g_{T2} and g_{T3} , their n is 3. For the OTA g_{T4} , its n is 6.

The CMFB loop is needed to stabilize the common mode (CM) output voltage. The reference CM voltage V_{CM} is set to be $V_{DD}/2$. Capacitors C_{1a} and C_{1b} with a value of 100 fF are used to perform frequency compensation, thus eliminating the risk of instability due to the CMFB loop. According to simulation results, at all the digital bit configurations, the OTA has an open loop gain > 30 dB, and its phase margin is > 60°. Table I shows the OTA's main components values.

TABLE I
MAIN COMPONENT VALUES OF THE OTA

W/L	$M_{1a,1b}$	$M_{2a,2b,3a,3b}$	$M_{4a,4b}$	$M_{5a,5b}$	$M_{6a,6b}$	M_{10a}	M_{10b}	$M_{15a,15b}$
$\mu\text{m}/\mu\text{m}$	60/0.36	12/0.36	100/0.18	60/0.18	9/0.36	5/0.36	10/0.36	6/0.36

3) Simulation of the 6th-Order LPF: To make the cut-off frequency of the 6th-order LPF $f_{c,6th}$ adjustable from 5 to 30 MHz and its pass-band gain $A_{LP,6th}$ tunable from 0 to 30 dB, the input current for each OTA I_{IN} is set to 10 μA and C_0 to 250 fF. The simulated settling time of the filter is of around 105 ns. Table II summarizes the simulated results of the OTA and the 6th-order LPF's $f_{c,6th}$ when the bit configuration is $b<2:0>$ from 000 to 111, and $b<5:3>$ equals 000.

B. BB Amplifier and Comparator

Fig. 4 shows the schematic of the proposed BB amplifier and comparator. According to simulation results, the BB amplifier has a 40-dB voltage gain with a -3 dB bandwidth of 40 MHz. The comparator compares the analogue BB signal with a DC

Fig. 3. Simplified schematic of the proposed OTA

Fig. 4. Schematic of the proposed BB amplifier and comparator

Fig. 5. Schematic of the proposed current bank

reference voltage set via its input pin V_{REF} . The BB amplifier and the comparator can be switched on and off together via the low enable signal $\bar{V}_{EN}$. The simulated turn-on time of the two blocks together is 50 ns. When $\bar{V}_{EN}=0$, the two parts draw in total $210 \mu A$ from a 1.8 V supply. Table III shows the main components values of this circuit.

TABLE II
SIMULATED RESULTS OF THE 6TH-ORDER LPF

b<2:0>	000	001	010	011	100	101	110	111
$I_{o,+} / \mu A$	10	15	20	25	31	37	42	47
$g_m / \mu S$	165	250	325	410	500	590	670	755
$I_{DD,6th} / mA$	0.665	0.81	0.935	1.06	1.21	1.345	1.46	1.591
$f_{c,6th} / MHz$	3.8	7	10	13.5	17.8	22.5	26	30.9

TABLE III
MAIN COMPONENT VALUES OF THE BB AMPLIFIER AND COMPARATOR

W/L	$M_{1a,1b}$	$M_{2a,2b}$	$M_{3a,3b}$	$M_{4a,4b}$	$M_{5a,5b}$	$M_{6a,6b}$	$M_{7a,7b,8a,8b}$
$\mu m / \mu m$	20/0.36	20/0.18	4/0.54	60/0.36	20/0.18	4.5/0.36	36/0.36

C. Current Bank

Fig. 4 shows the schematic of the proposed current bank. This current bank provides 12 outputs $I_{OUT}<11:0>$ for the LPF, BB amplifier and comparator. Each output $I_{OUT}<x>$ sources $10 \mu A$. The current bank can be switched on and off via $\bar{V}_{EN}$. The simulated settling time of the output current $I_{OUT}<x>$ is 78 ns. When $\bar{V}_{EN}=0$, the current bank excluding the currents on the output paths consumes $37.5 \mu A$ from a 1.8 V supply. Table IV shows the main components values of the designed current bank.

TABLE IV
MAIN COMPONENT VALUES OF THE CURRENT BANK

W/L	$M_{1a,1b}$	$M_{2a,2b}$	$M_{3a,3b}$	$M_{4a,4b}$	$M_{5a,5b}$	M_{6a}	$M_{6b,6c}$	M_{7a}	M_{7b}
$\mu m / \mu m$	4/0.18	2/20	4/0.18	2/0.72	4/1	1/3.6	41/0.18	4/0.18	16/0.18

D. Digital Control

To enable the external digital control of the circuit, a customized slave SPI is implemented on the chip. To communicate with this slave SPI, an STM based micro-controller PCB is developed. Moreover, a Python based graphical user interface is developed to perform experiments efficiently.

III. EXPERIMENTAL RESULTS

A die photograph of the completely integrated frequency detector is shown in Fig. 6. The BB chain excluding the digital SPI occupies a silicon area of 0.3 mm^2 . For test purposes, the chip is wire-bonded on a PCB. Fig. 6 and Fig. 7 show the photograph and the schematic of this PCB, respectively. On the PCB, the positive input of the LPF, which is connected to a 50Ω resistor, is used as the RF signal input of the circuit. The negative input of the LPF is biased to the common mode voltage of the circuit $V_{DD}/2$. In addition, a broadband fully differential buffer (Linear LTC6417) with a -3 dB small signal bandwidth of 1.6 GHz is used as the differential 50Ω output buffer for the LPF. A broadband balun is used to convert the filter's differential output to a single-ended output.

A. Characterization of the 6th-Order LPF

A vector analyzer Rohde&Schwarz ZVL-6 was used to obtain the S-parameter of the circuit. Fig. 8 shows the 8 fre-

quency responses corresponding to the filter bit configuration of $b<5:3> = 000$ and $b<2:0>$ from 000 to 111. Here we can first see that the cut-off frequency $f_{c,6th}$ is changed accordingly to the corresponding bit configuration. We can also tell that the LPF achieves a transition steepness of -120 dB/decade . However, the filter's passband gain $A_{LP,6th}$ decreases when the cut-off frequency f_c becomes higher, which is a result of process variations. To compensate the loss of $A_{LP,6th}$, the higher 3 bits of the LPF, $b<5:3>$, are tuned. Fig. 9 shows the 8 frequency responses after adjustment of the filter's $A_{LP,6th}$. Here we can see that the reduction of the passband gain can be successfully compensated, and the specified filter characteristics are satisfied. In the following tests, we set the filter's bit configuration $b<5:0>$ to 011011, which corresponds to a DC current consumption of around 1.34 mA at 1.8 V .

To characterize the filter's transient behavior, we used a waveform generator Keysight 33600A with two channels, that generate, respectively, a 10 MHz and 20 mV_{PP} sine wave signal for the LPF's input as well as a 10 KHz low enable signal for the LPF's $\bar{V}_{EN}$. The settling time is defined as the time slot the LPF takes from the start of its power-up signal until its output signal reaches 90% of its steady state amplitude. Fig. 10 shows the measured filter's output signal with its

Fig. 6. Chip photograph of the fabricated frequency detector

Fig. 7. Schematic of the PCB with the chip under test

Fig. 8. Measured filter frequency responses without $A_{LP,6th}$ compensation

Fig. 9. Measured frequency responses with $A_{LP,6th}$ compensation

Fig. 10. Measured filter output signal with $\bar{V}_{EN}$ being applied

low enable signal $\bar{V}_{EN}$. It can be seen that $\bar{V}_{EN}$ starts to drop at $0.5 \mu s$, and the LPF settles at around $0.614 \mu s$. Thus, the directly observed settling time is $140 ns$. This time slot, however, includes the delay time caused by the on-board output buffer, the transmission line on the PCB, the balun and the cables, which according to our characterization is approximated to $30 ns$. Thus, the measured settling time is $110 ns$ which matches the simulated value of $105 ns$ very well.

The performance of the LPF and comparison with prior works are summarized in Table V. Here we can see that the fabricated 6th-order LPF achieves the lowest DC power consumption per pole among the state-of-the-art LPFs. Moreover, it demonstrates a fast switchable functionality that the other LPFs do not have. The switch-on time of $110 ns$ is very low. Thanks to the low power consumption and fast switchability, our proposed LPF is an ideal candidate for ultra low power and especially duty-cycled applications, such as, wake-up receivers.

TABLE V
COMPARISON WITH STATE-OF-THE-ART LPFS

	This work	[13]	[14]	[15]	[16]
Technology	180 nm CMOS	180 nm CMOS	180 nm CMOS	90 nm CMOS	65 nm CMOS
Topology	TT g_m -C	Active RC	Sallen Key	g_m -C	Active RC
Filter order	6	4	4	6	2/4/6/8
Settling time (ns)	110	-	-	-	-
V_{DD} (V)	1.8	1.8	1.8	1.0	1.2
P_{DC} (mW)	2.41	12.4	3.2	4.35	1.72 - 9.6
P_{DC} per pole (mW)	0.4	3.1	0.8	0.725	0.86 - 1.2
f_c (MHz)	10	22.5	28.8	8.1 - 13.5	0.2 - 20
Passband gain (dB)	0.5 - 4	0.5	-3	0 - 20	2
Size (mm ²)	0.3	0.357	0.12	0.239	0.8

Fig. 11. Time trace of the PRBS, the output of the LPF and comparator.

B. Characterization of the BB Chain

A bit error rate (BER) test was performed to quantitatively characterize the functionality of the analogue BB chain. In order to model the output of an MTJ device, we used the PRBS function of a waveform generator Keysight 33600A to generate a pseudo-random bit stream (PRBS). The data rate of the PRBS was set to 5 Mbps. The reference voltage of the comparator V_{REF} was set to 0.7 V via the micro-controller. An oscilloscope Rohde&Schwarz RTO 1044 was used to sample the PRBS, the output of the LPF and the comparator. The sampled bit length of the PRBS was set to 10^6 . V_{PP} of the PRBS was swept from 3.5 mV to 1.5 mV with a step of 0.5 mV. Fig. 11 shows a $40 \mu s$ time trace of the PRBS, the output of the LPF and comparator, when the PRBS was at a data rate of 5 Mbps with V_{PP} equal to 2 mV. It can be seen, for the observed time slot, the BB chain has successfully interpreted the input BB signal to the corresponding digital bit flow.

Moreover, a Matlab script was developed to compare the data between the PRBS and the output of the comparator and calculate the BER of each test case. Table VI summarizes the BER of the BB chain at different V_{PP} of the input BB signal. It can be concluded that if the BB signal converted by the MTJ has a V_{PP} greater than 2.5 mV or even 2.0 mV, the MTJ based RF signal detector can work properly.

TABLE VI
MEASURED BER OF THE BB CHAIN AT 5 MBPS

V_{pp}	3.5 mV	3.0 mV	2.5 mV	2.0 mV	1.5 mV
BER	$< 10^{-6}$	$< 10^{-6}$	$< 10^{-6}$	$1.3 \cdot 10^{-5}$	$2.5 \cdot 10^{-4}$

IV. CONCLUSION

This work has presented an analogue BB chain for an MTJ based RF signal detector fully integrated in a hybrid CMOS-MTJ technology. The BB chain contains a fully-differential 6th-order g_m -C based LPF, a BB amplifier, a comparator, a current bank and a digital SPI.

The fabricated LPF with a cut-off frequency of 10 MHz achieves a very low DC power consumption of 2.41 mW. Its DC power per pole of 0.4 mW is the lowest among the state-of-the-art LPFs. Moreover, the LPF can be switched on and

off very fast within 110 ns. The fast switch-ability, remarkably low power consumption and a competitively small silicon area allow this LPF to improve the state of the art.

In addition, the complete analogue BB chain is able to transform a 2.5 mV_{PP}, 5 Mbps BB signal into digital data with a BER fewer than 10^{-6} . The BB chain consumes a DC current of 2.85 mA from a 1.8 V supply including all bias circuits. To achieve power efficiency, the BB chain is able to operate under the duty-cycling mode. The switch-on time of the BB chain is within 200 ns.

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