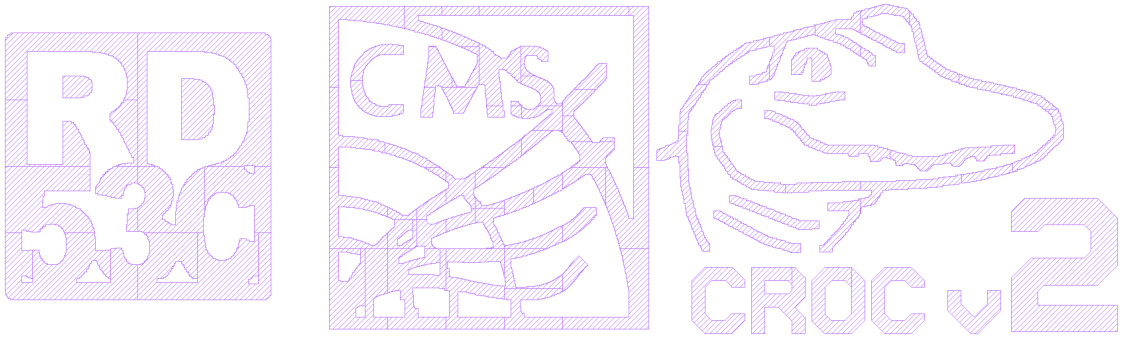
CERN-RD53-PUB-xxxx
Version 1.92, September 18, 2024

The RD53C-CMS Pixel Readout Chip Manual

ABSTRACT: Manual for the RD53C design in the CMS chip implementation.



Contents

1. Overview

This manual provides a technical description of the RD53C chip design and operation adequate for simulation, testing, debugging and DAQ development. A basic familiarity with pixel systems and readout chips is assumed. A more general introduction explaining the basic functions and principles can be found in [6].

The production readout chips for the ATLAS and CMS HL-LHC pixel detectors are two separate instances of a common design framework called RD53C. The main differences between ATLAS and CMS are the size of the pixel matrix and the pixel analog front end. There are other differences partly stemming for the sequential fabrication: RD53C-ATLAS in spring and RD53C-CMS in fall of 2023. RD53C-CMS has a few minor added features relative to RD53C-ATLAS. For convenience this manual is compiled in two separate versions, RD53C-ATLAS and RD53C-CMS. This version is for the RD53C-CMS chip, designated by CMS as CROC-V2. Both manual versions use the same revision number as most of the elements are common. RD53C is an evolution of the RD53B framework [**?**] and RD53B-ATLAS and RD53B-CMS pre-production chips. The requirements were defined by the experiments for RD53B [2] and have not changed for RD53C.

RD53C is a pixel readout chip framework that can be instantiated into different size physical chips. The design work and much of the verification are largely independent of the final instantiated size. RD53C consists of a *pixel matrix* and a *chip bottom*. The pixel matrix is built up of identical 8 by 8 pixel *cores* stepped and repeated in columns and rows. A core is physically 400 μm by 400 μm. The selected numbers of core columns and rows determine the chip size. The chip bottom contains all the system functionality and should be viewed as a fixed element that does not depend on matrix size. A physical chip, therefore, cannot be *narrower* than 20 mm (50 cores), because that is the width of the unique wire bonding pad frame in the chip bottom, but it can be wider. The *height* (number of core rows) is not constrained by the chip bottom, but is limited to a maximum of 50 by power and bias distribution as well as readout timing. This high level organization concept is shown in Fig. 1. The instantiated dimensions are detailed in Sec. 2.

The core contains 64 pixel front ends organized in 16 identical so-called *analog islands* with 4 fronts ends each, which are embedded in a flat digital synthesized "sea" as shown in Fig. 2. The analog front end and island design are described in Sec. 5. The digital core design is described in Sec. 7. The pixel matrix is produced by stepping and repeating identical cores, which also takes care of the distribution of analog biases, as described in Sec 5.1.

The chip bottom contains all system functionality and the wire bond pads. RD53C is a system-on-chip including power management, sophisticated digital communication, sensing and monitoring. An overview of the basic operation, including a description of the reset scheme, is given in Sec 3. Sec 3 also serves as an introduction to the more detailed content of other sections. All tabular information, including pinout and configuration register values, is collected in the Reference section (Sec 16).

Power management, including design of the Shunt-LDO regulators are covered in Sec 4. The command and control interface (how one talks to the chip) and the configuration are covered in Sec 8. The data output (what comes out of the chip), including special (non-hit data) and the aggregation of data from multiple chips, are described in Sec. 10. The sensing and monitoring functions are described in Sec 12. Test features and miscellaneous functions are covered in Sec 13.

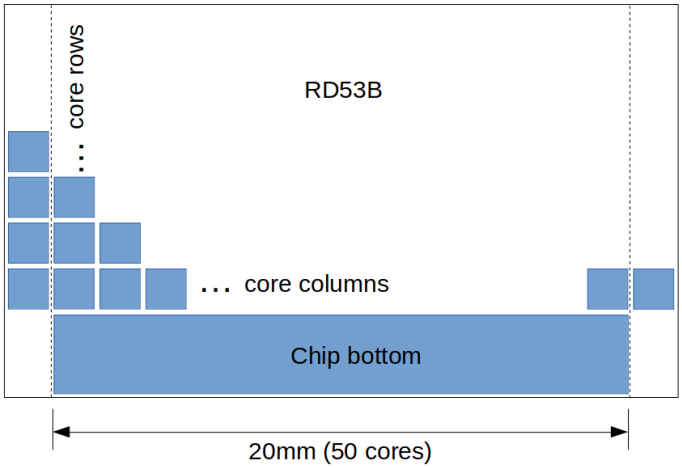


Figure 1: Conceptual depiction of RD53C framework, with a matrix composed of 50 or more columns by up to 50 rows of identical cores, and a fixed chip bottom. The dashed lines indicate the minimum width of 50 cores. Core number 0,0 is at the top left of the figure, while the highest column, row numbered core is at the bottom right.

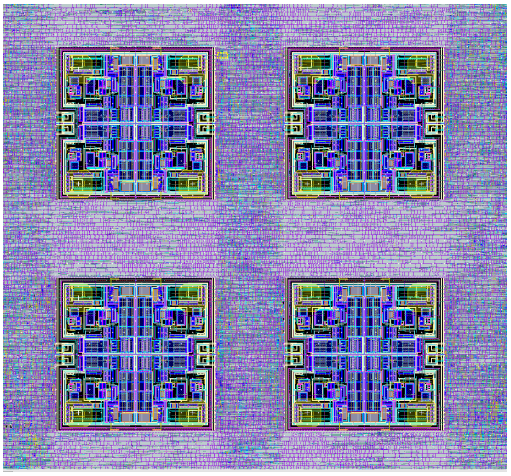


Figure 2: Layout view of analog islands within synthesized logic. Four complete islands can be seen in the center of the figure. One core contains four by four analog islands.

The designs of the bump bond and wire bond pads are covered in Sec. 2.1 and 2.2. RD53C only has wire bond pads along the bottom edge, to make it 3-side abuttable.

Parameter	ATLAS	CMS
Pixel bump pitch	50 μm $\times$ 50 μm	
pixel rows (H)	384	336
pixel columns (W)	400	432
core rows	48	42
core columns	50	54
Chip width (including seal ring)	20.050 mm	21.654 mm
Chip height (including seal ring)	21.0213 mm	18.622 mm

Table 1: Size of ATLAS and CMS chips in cores and outline measured from outside edge of seal ring.

2. Dimensions, Floorplan and Pads

RD53C uses a 9 metal layer stack, consisting of 7 thin, 1 thick and 1 ultra-thick metal layers. In addition, the 28K AP layer is also used for power lines distribution. In Fig. 3 the layout and functional view of RD53C floorplan are shown. The sensitive area of the chip is placed at the top of the chip and is arranged as a matrix of pixel bump pads on 50 μm $\times$ 50 μm pitch according to Table 1. The peripheral circuitry is placed at the bottom of the chip and contains all global analog and digital circuitry needed to bias, configure, monitor and readout the chip. The wire bonding pads are organized as a single row at the bottom chip edge and are separated from the first row of bumps by 1.7 mm in order to allow for wire bonding after sensor flip-chip (Sec. 2.2).

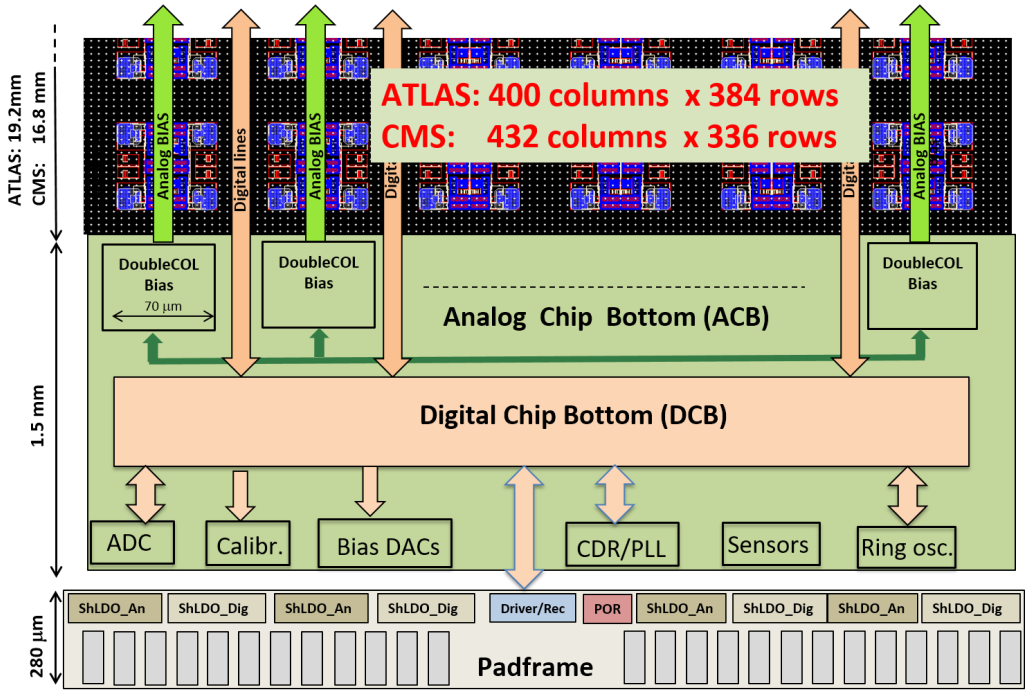


Figure 3: RD53C floorplan, functional view.

In the chip periphery, all the analog building blocks are grouped in a macroblock called Ana-

log Chip Bottom (ACB), which is fully assembled and characterized in an analog environment. The ACB block is surrounded by a synthesized block, called Digital Chip Bottom (DCB), which implements the Input, Output and Configuration digital logic.

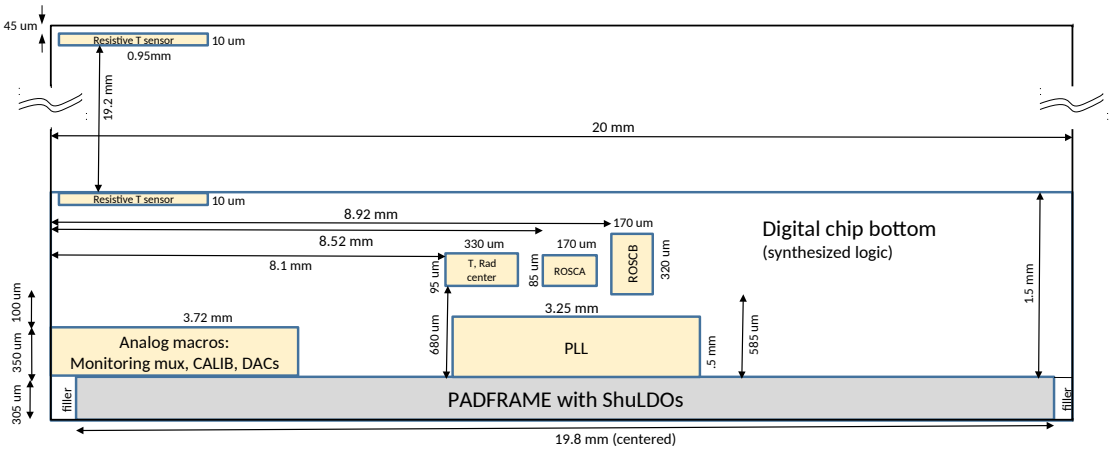


Figure 4: Size and location of elements in the CMS chip bottom and top (Not to scale). The outline is the chip seal ring (tightest possible diced edge).

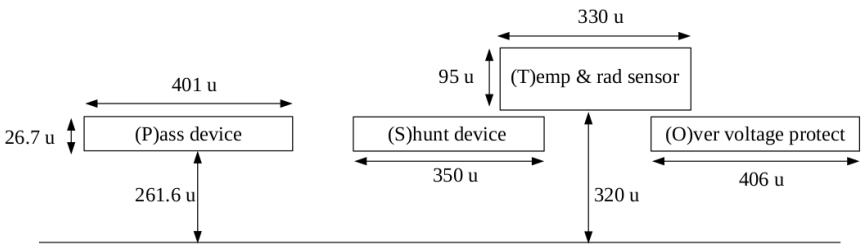


Figure 5: Size and vertical position of power devices relative to the chip seal ring (tightest possible diced edge). Horizontal placement is given in Table 2

Device	AO	AS	AP	DP	DS	DO	AO	AS	AP	DP	DS	DO
Left edge (um)	954	1371	1715	2734	3139	3480	6154	6571	6715	7934	8339	8680
Device									AT	DT		
Left edge (um)									7125	7825		
Device	AO	AS	AP	DP	DS	DO	AO	AS	AP	DP	DS	DO
Left edge (um)	11354	11771	12115	13134	13539	13880	16654	17071	17415	18434	18839	19180

Table 2: Companion table to Fig. 5 showing the position of the left edge of power devices in ATLAS chip relative to the outside edge of the seal ring: A=Analog, D=Digital, O=Over voltage protection, P=Pass device, S=Shunt device, T= Temperature and radiation sensor. The devices are arranged in four groups (delimited by double lines) as can be seen in Fig. 8.

2.1 Bump Bond Pads

The bump bonds pads are defined by a regular pattern of openings in the passivation as shown in Fig. 6 (left). The alignment of aluminum metal shape under each passivation opening can vary by up to 1 μm from pixel to pixel, but as the shapes are bigger than the opening there is always exposed aluminum for the entire pad. The bump bond pads do not have ESD protection. The passivation opening is square with 45 degree corners, which will appear rounded in the as-built chip. Fig. 6 (right) shows the expected height profile across the center of a bump pad as derived from the metal stack. Aluminum metal is exposed in the 12 μm passivation opening and extends below the passivation beyond the opening, resulting in a passivation ridge surrounding the opening, as shown. The exposed metal may not be completely flat: it can have depressions less than 1 μm deep due to vias below. The figure shows such a depression.

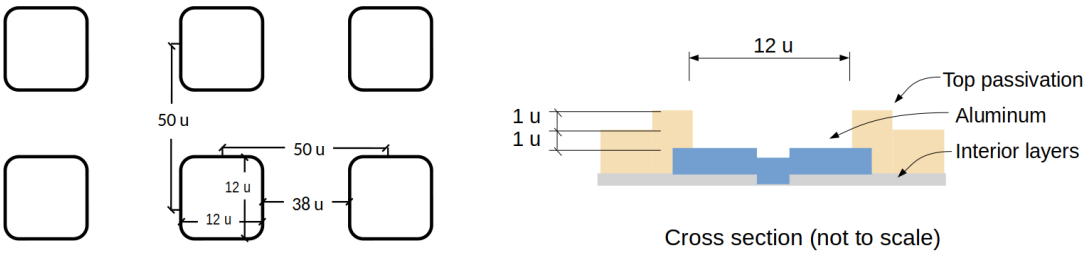


Figure 6: Bump bond pad dimensions. Matrix layout on the left and cross section on the right.

2.2 Wire Bond Pads and Alignment Marks

The wire bond pads are along the chip bottom on a 100 μm pitch. The pad area is large enough to meet production requirements (Fig. **??**). There are 198 pads, 4 of which are not used and not connected to any internal net. The location of these unused pads was chosen to eliminate wire bonding tool interference at the edges of fanout regions. Most pads are for power and ground and are grouped strategically for PCB/module layout as shown in Fig. 8. The detailed pinout is given in Sec. 16.1.

The wire bond pads have visible numbering on the chip (the numbers label the pads to their right), and are flanked by alignment marks, as can be seen in Fig. 7.

The RD53C chip has internally four separate power domains:

- Analog: VDDA, GNDA
- Digital: VDDD, GNDD
- PLL (PLL/CDR + CMD_IN LVDS receiver): VDD_PLL, GND_PLL
- CML (serializer + cable driver): VDD_CML, GND_CML

The local ESD devices connect to both power and ground rails or to the ground rail only in case of over-voltage tolerant pads (OVT). OVT pads are used where the input voltage could potentially exceed the local power rail (see pad listing in Sec. 16.1). In a typical environment, all ground rails are wire-bonded to the same system ground, which enables ESD paths between the (otherwise

isolated) power domains. However, during assembly or wafer probing, a common external ground rail might not be established yet. To account for this, a common ESD bus (VSS) has been used to connect the different ground rails via on-chip anti-parallel diodes to create a safe ESD path between power domains at all times. This net (VSS, also used for connecting the global substrate VSUB) **should be wire-bonded first** (pads 9, 91, and 196), then all remaining ground pads, and finally the rest of the pads.

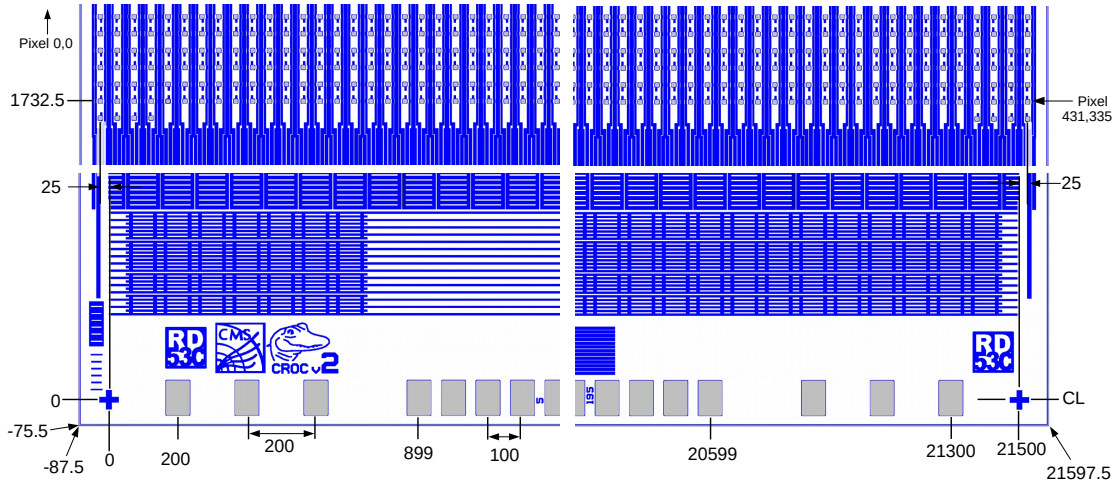


Figure 7: Detail of CMS chip dimensions. The location of the first and last pixel bump bonds on the matrix is also indicated. There are 4 bump bond pads below the full matrix on each of left and right sides to contact sensor bias/guard rings.

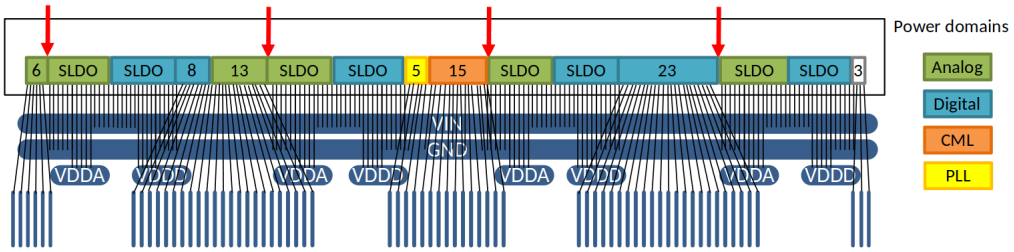


Figure 8: Organization of wire bond pad frame and generic bonding scheme. All wire bonds are shown, including connections for testing (not used on detector modules). The number of fanned-out signal bonds is written in each box, while the power supply bonds run parallel (not fanned out). The red arrows indicate the four unused pads.

3. Basic Operation and Reset (Intro and Quick Start Guide)

This section walks through the steps for basic, beginner level operation of a single chip on the bench. It also describes how the chip is reset- a critical point for correct start-up. Advanced users will often do things differently than stated in this introductory section, and so the more detailed sections of the document are referenced as appropriate. Thus, this section can also be used as a guide to the rest of the document. Each item in this section is just one choice out of several possible connections and configuration values (recommended for initial operation). Whenever registers or pins are mentioned they can be found in the reference section (16).

3.1 Chip Startup

The startup sequence is power, clock, communication, configuration, operation.

Power Typical bench testing will use the LDO powering option, in which the internal regulators are used as classic linear regulators fed from a constant voltage power supply, rather than serial power regulators fed from a constant current supply as they will be used in the experiment. LDO powering is more convenient for single chip testing (LDO stands for Low Drop Out voltage). A single chip card will contain jumpers to select LDO mode. A single power supply will be connected to all the chip's V_IN pins, while the shunt mode controller voltage, VDD_SHUNT, will be disconnected, which is all that is needed to disable shunt mode. For details see Sec. 4, which also describes the serial power and direct power configurations. External components should be set to their nominal values (Table 30). A power supply current limit of 2 A (half analog and half digital) will be typical.

LDO mode allows to view the internal current consumption. (VINA and VIND can be connected individually to monitor currents in analog and digital domains). The supply voltage should be a minimum of 1.4 V and never more than 2.0 V. 1.6 V should be a typical setting to have comfortable margin for cable voltage drops. When power is turned on, the current consumption will be determined by the default configuration, which is low (normal) power for the ATLAS (CMS) chip. Typical current consumption is given in Table 3. The VDDA and VDDD regulator outputs, which connect to external decoupling capacitors, should produce approximately 1.2 V, which is the default setting (one should verify that this is the case when first testing a chip). Both the regulated voltages and current consumption will be affected by the main reference current (Iref), which has a nominal value of 20 μA and can be trimmed with wire bonds (or jumpers on a single chip test card) if needed (Sec. 4.2). Without any IREF wire bonds (or jumpers on a single chip test card) the current reference will be at its maximum value, significantly more than 20 μA, while default and recommended configuration settings assume 20 μA. Since all internal biases are derived from this current reference, all bias currents will be high in this case. A very quick and dirty solution to this is to wire bond (or load the jumper on a single chip card) the most significant bit, which will trim to the middle of the trimming range.

Clock The PLL Clock Data Recovery circuit will become active as soon as it has power and will produce clock edges on all the internal clocks even in the absence of any external command input. But these will have arbitrary frequency and phase. This arbitrary clock is useful as a diagnostic: it will drive the data output stage (all four CML outputs) and produce a "heartbeat" idle pattern that

	ATLAS (mA)			CMS (mA)		
Status	**Analog**	**Digital**	**Total**	**Analog**	**Digital**	**Total**
Power only (I/O unplugged)	160	185	345	650	400	850
Communication up	165	270	435	650	430	1080
Configured for testing	700	740	1440	800	740	1540

Table 3: Typical current consumption for single chip bench test operation, assuming a nominal (20 μA) reference current (Iref). Unconnected Iref pins/jumpers can result in 15% higher values. The total column is what should be observed when using a single power supply, as recommended.

can be observed to confirm that the chip is alive and the data connections present. But this arbitrary clock is not useful for operation. For that one needs a known frequency and phase clock that is obtained by locking to the incoming command bitstream.

The initialization and reset procedures needed to establish a proper clock and communication will be carried out automatically by the DAQ without user intervention, but they are described here to provide a basic introduction to how the chip operates and allow troubleshooting.

The reset organization is described in Sec. 3.2. Regardless of the state of the command input during power up, after power is stable, communication must be initialized by first "idling" the command line to a nominal bitrate of 1 Mbps[1] for at least 10 μs, and then supplying a 160 Mbps clock pattern (80 MHz effective clock frequency) for at least 1 ms. The clock pattern is an uninterrupted stream of PLL_LOCK symbols (Sec. 8.2). This is equivalent to a No Operation (NOOP) command in many processors, and can be used as filler when no other commands must be sent, but it will be referred to as PLL_LOCK or PLLlock in RD53C.

This "idling" of the command line is the main hard reset mechanism for RD53C. It can be done at any time to recover the chip from a bad state without a need to power cycle. It should be thought of as the equivalent of power cycling hard reset, so it is a reset tool of last resort. After the command idle reset, the PLL will enter lock mode, and supplying a clock pattern is critical for it to lock to the correct frequency. The locking of the PLL can be verified with an optional diagnostic output (see below), but during detector operation this diagnostic will not be available and there will be no external indication that the PLL has locked. It will therefore be necessary to hold the clock pattern long enough to leave no doubt that there has been enough time to lock (1 ms). Further details are given in Sec 14.

Optional Diagnostics During bench testing it is possible to access a variety of test outputs. The chip has one CMOS and four LVDS general purpose outputs that can show a selection of internal signals (see Sec. 16.4). By default these carry the following information:

CMOS: gpo_ch_sync_lock: 1 if the ChannelSync is locked, 0 when it is unlocked (see below).

LVDS_0: CMD_raw: repeater of the chip command serial input. A buffered version of what the chip receives.

LVDS_1: cdr_cmd_data_predel: sampled input command pattern before applying any delay (should be very similar to the above).

[1] A DC level- low or high- instead of 1 Mbps will also work for initiating the reset, but is not advised for A/C coupled command lines.

LVDS_2: PorResetB: output of Power On Reset circuit (active low) in case it is needed (not normally used in RD53C).

LVDS_3: gpo_cdr_lock_status: PLL Lock signal. 1 if Locked, 0 otherwise (see clock above).

Communication Now the all clocks will be at the correct frequency and the gpo_cdr_lock_status shown above should be high. But the chip is not yet ready to understand commands, because the alignment of the incoming command frames has not yet been established (gpo_ch_sync_lock should still be low). This alignment is done by a circuit called the channel synchronizer, that recognizes unique bit patterns called sync symbols (all command symbols are described in Sec. 8). Therefore, after the clock pattern, the sending of sync symbols will be enabled (once again, the DAQ system will do this automatically). One can send a constant string of sync symbols or simply enable automatic insertion of one sync every N frames (where N is set by configuration, default 32). So either (sync, sync, sync,...) or (PLLlock, PLLlock, PLLlock, sync, PLLlock, PLLlock,...)- it makes no difference. The important thing is to send a large number of sync symbols (exceeding a minimum number set by configuration, default 16) before sending any commands. When the channel synchronizer locks, the gpo_ch_sync_lock signal will go high on the general purpose CMOS output. Commands will not be accepted (so the chip configuration cannot be changed) unless this lock signal is high. Again, the DAQ will normally ensure the correctness of this sequence with no need to look at the diagnostic signals.

The CMS chip has a reduced power default configuration. The chip starts with a power consumption a bit lower than nominal: the analog FE are set to absorb 4 μA instead of the nominal 5 μA, the global threshold is set to a very high value and the comparators are set in such a way to prevent anomalous currents at startup. Moreover, only 1/3 of pixels receive the clock.

In general a new configuration will have to be loaded for most single chip testing. Test setups will include a baseline configuration suitable for most tests (which may also be called default in test setup documentation, should not be confused with the internal chip default configuration).

3.1.1 Default and User Configurations

When the chip starts up and is reset, (A) the global configuration will be supplied by internal hard-wired default values and (B) the user programmable configuration registers will be automatically loaded with those same default values. This complex scheme of having two configurations (hard-wired and programmed) is needed to ensure that the default configuration is present immediately upon power-up or upon CMD idle reset, regardless of the presence of a clock, or of the time it takes to load the programmable configuration registers. Which of the two configurations controls the chip is determined by multiplexers associated with each register, as indicated in Fig. 9. At start-up, the hard-wired default global configuration will be controlling the chip. The same mechanism is implemented for the pixel configuration, but unlike the pixel matrix, where the programmed configuration has no reset at all, the global configuration has both the MUX and a synchronous reset, so that whenever the MUX selects the hard-wired configuration, the programmed values will soon (when clocks arrive) be reset to equal the hard-wired values.

In order to use a programmed configuration different from the default, control of the chip must first be switched over from the hard wired to the programmable configuration. The configuration selection is controlled by both the CMD idle reset signal and a logic comparator that

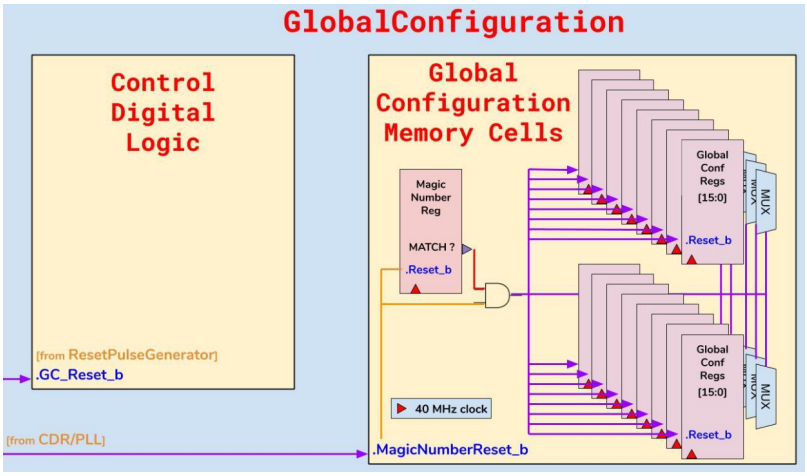


Figure 9: Configuration selection and reset.

compares the value stored in a pair of special configuration registers (32 bits total) to a hard-wired key code or "magic number" (labeled MagicNumberReg in the figure). When the stored value does not match the key, the hard-wired configuration is selected. Since at power-up the registers will contain something arbitrary, and will be reset to zero as soon as clock edges are present, the hard-wired configuration will be selected. To switch over control of the chip to the programmed configuration, the user must write the key into the magic number registers (the key code is Hex AC75 in GCR_DEFAULT_CONFIG and Hex 538A in GCR_DEFAULT_CONFIG_B, as can be seen in Table 22). Since at start-up the programmable registers will have been initialized to the default configuration, when the magic number is written and the control of the chip switches over, nothing should actually change, because each register is switched from the hard-wired default to the same default in a its programmable register. This is important to avoid a sudden current jump since all registers switch over at the same time. Now each programmable register can be written to the desired value, one at a time.

To guard against SEU, in addition to being triple redundant (as are all global configuration registers), any permutation of the key codes with one bit flipped will also select the programmed configuration. In addition to a possible bit flip in the key code, an SEE could put a glitch in the MUX control level and that will cause the active configuration bits to switch between hard-wired and programmed for the duration of the glitch. This is relatively benign and not persistent: after the glitch everything will be in the original state and there is no corruption of the stored configuration. Most configuration bits control DACs, which have a slow response time and will therefore not propagate a glitch in their control bits to their output analog level.

3.2 Reset

The driving requirements of the RD53C reset scheme are:

- Avoiding introduction of Single Event Effect vulnerability. This led to having reset capability only on circuits that absolutely need it, and to use only synchronous reset for them. A synchronous reset signal is a logic level that is sampled locally every clock edge. Spurious

glitches on this reset signal have no effect (in contrast to an asynchronous reset, for which a transition produces a reset regardless of clock).

- Need for a default configuration present immediately upon power up. This is done without the use of a power-on reset, as this would require an asynchronous reset on the global configuration registers. The default configuration is not stored in registers, but hard-wired and selected by a 2-to-1 multiplexer (Sec. 3.1.1)

- The ability to reset a chip (or a subcircuit within) without cycling the power, which would require tuning off and on an entire serial chain. This is accomplished with activity detection on the command input (Sec. 3.2.1)

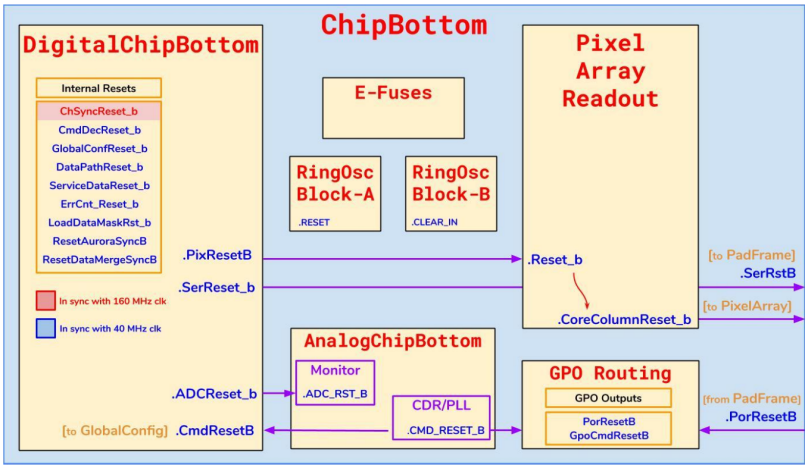


Figure 10: Block diagram of reset signals in the RD53C chip as described in the text.

The overall reset organization is shown in Fig. 10. All signal use negative logic: low means reset. There is a power-on reset generation circuit in the chip, inherited from RD53A, but the output of this circuit is not used to reset anything in RD53C and is only sent to the general purpose output multiplexer so that is available for external routing. The only asynchronous reset signal that is used in RD53C is the command activity detector (Sec. 3.2.1), labeled .CMD_RESET_B in the figure. This signal performs 3 functions: (1) it resets the PLL circuit that recovers the clock, (2) it selects the default configuration (Sec. 3.1.1), and (3) it is used (after synchronization) to actuate all the synchronous resets in the digital chip bottom.

All digital blocks have synchronous resets. These can be individually actuated at any time using the Global Pulse command, in addition to the actuation by the synchronized .CMD_RESET_B signal in (3) above. The organization of the digital block resets is shown in Fig. 11. The global configuration registers are explained in Sec. 3.1.1. The logic to write and read global configuration has its own synchronous reset, labeled .GlobalConfReset_B in the figure.

3.2.1 Command Activity Detector

The purpose of this circuit is to provide a “hard reset” mechanism for PLL/CDR block that recovers the clock from the input command stream and controls internal resets. The command activity

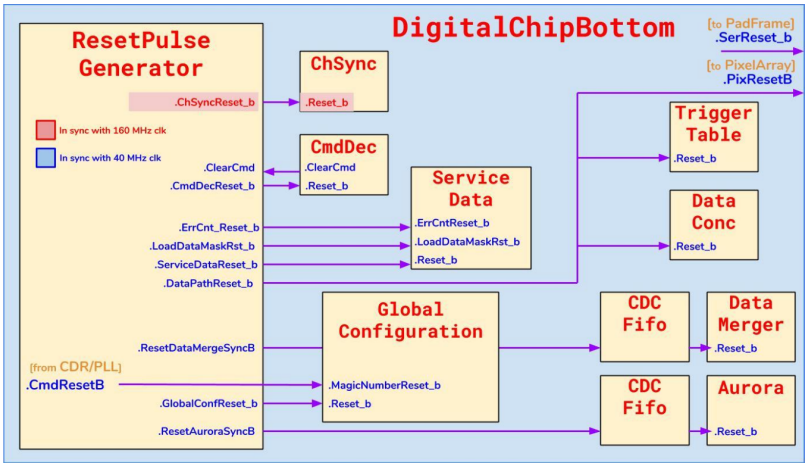


Figure 11: Block diagram of reset signals in the RD53C digital bottom as described in the text.

detector measures the rate of transitions in the incoming command signal. A positive edge rate below a nominal 10 MHz causes a reset to be asserted, while a higher frequency removes the reset. This nominal 10 MHz threshold has a significant uncertainty, with process, voltage, and temperature dependence. Thus, an edge rate «10 MHz (called idling) should be provided to guarantee reset, while normal command activity has a positive rate always between 30 MHz and 80 MHz. The circuit bandwidth is low enough that it takes of order 2 μs after command line idling for the reset to be asserted. It will take of order 0.5 μs to release the reset once the command line is returned to normal.

The activity detector is part of the PLL/CDR block. It directly resets the PLL, which means it puts it back into frequency lock mode.

This is the main hard reset mechanism of the RD53C chip, conceptually equivalent to cycling the power in a typical system. This is necessary because in a serial power chain, cycling the power is truly an action of last resort that should never be needed.

4. Power and References

RD53C is designed for operation in a serial powered system, where multiple chips are powered in parallel within a module, and multiple modules are connected in series. All circuits needed for such operation are built into the chip, such that only passive external components are needed to implement serial power chains. The foundation of this system a custom constant current regulator called Shunt LDO (SLDO). In addition to the SLDO proper, RD53C contains auxiliary circuits including voltage/current references with automatic start-up (Sec. 4.2), "under-shunt" transient protection analogous to a current limit for constant voltage supplies (Sec. 4.4), overvoltage protection (Sec. 4.5), and a low current operation mode for detector tests with limited cooling (Sec. 4.3).

4.1 Shunt LDO Regulator

The Shunt LDO regulator (SLDO) regulator is a combination of a linear Low Drop-Out voltage regulator (LDO) and a shunt element. The goal is to provide constant current operation with multiple devices connected in parallel (which is not possible for conventional shunt regulators). The circuit was invented as part of the FE-I4 chip development [4], but the design has evolved significantly towards the final implementation in RD53C.

The basic principle of operation of the SLDO circuit can be explained using Fig. 12. A conventional LDO voltage regulator is used to power the main load, L, as usual, plus an internal load, L_s, in parallel. This internal load (referred to as the shunt element, hence L_s) is actively controlled to achieve the desired behavior at the input, no matter what the main load L does. To first order, the desired behavior is $I_{in} = I_L + I_s = constant$.

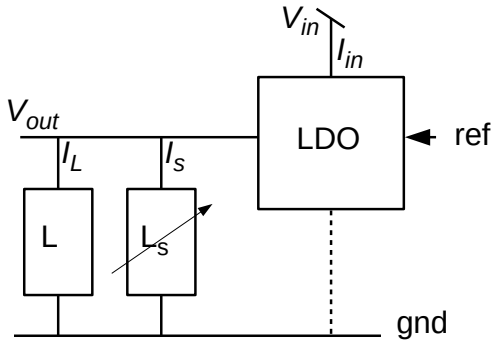


Figure 12: Concept of SLDO operation as a linear regulator (LDO) powering an main load L and a variable internal shunt load L_s.

The real needed behavior for serial power operation is more complex in order to achieve efficient current sharing among parallel chips and is given by Eq. 4.1.

$$I_{in} = I_L + I_s = \frac{V_{in} - V_0}{R_{\text{eff}}} \qquad [V_{in} > V_0] \tag{4.1}$$

where V_0 is a constant but programmable offset needed for high efficiency and R_{eff}, also user programmable, gives an ohmic behavior necessary to share current evenly among parallel devices. A diagram of the desired behavior is shown in Fig. 13.

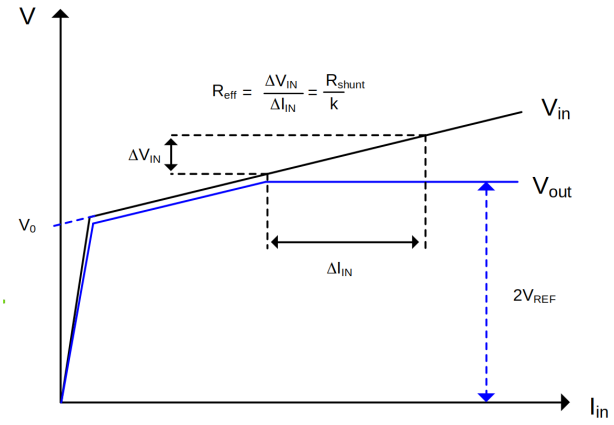


Figure 13: Desired current vs. voltage characteristics for SLDO. The unregulated input voltage and regulated output voltage are shown. Indicated values are discussed in the text.

The simplified circuit schematic of the RD53C SLDO is shown in Fig. 14. The red part of the circuit is a classic LDO regulator with pass device M1. The rest of the circuit can be disabled in order to operate in pure LDO mode, which is useful for testing individual chips and for observing the current consumption. The main load L (external to the SLDO) is not shown- it is the chip itself. Device M4 is the internal load, L_s, of Fig. 12 and the rest of the black circuitry is the active control. This control ensures that the current in the pass device M1 is equal to the the current through R_{shunt} (I_{ctrl} for control current) times the scale factor K, which has a design value of 1000. R_{shunt} is an external resistor to allow the user precise control R_{eff}. It can be seen that $R_{\text{eff}} = R_{\text{shunt}}/1000$. Finally, the blue circuit provides the offset V_0. This is controlled by a reference voltage labeled V_{ofs}. User control of V_{ofs} is described in Sec. 4.3.

The SLDO circuit is designed to be compatible with 2 V input voltage. All transistors are cascoded in order to always have more than two transistors between voltage supply and ground, with supply voltage distributed across several transistors. Device voltage limits checks in static and dynamic simulations show that no transistor sees more than 1.32 V across any two terminals, even during transients. The one exception is the pass device M1, where cascoding to protect against over-voltage would cause higher drop-out voltage and therefore higher power consumption. This lack of cascoding of device M1 leads a lower limit to the undershunt protection range equal to Vref (Sec. 4.4). The SLDO circuit also uses a Low-ESR output capacitor compensation scheme, such that careful control of the external component equivalent series resistance (ESR) is not necessary.

4.2 References and Startup

For serial chain operation the SLDO must become active immediately upon current flow, before communication is possible. Once operational, it must work with high efficiency and uniformity among chips in the chain. Furthermore, startup must work reliably over a wide temperature range,

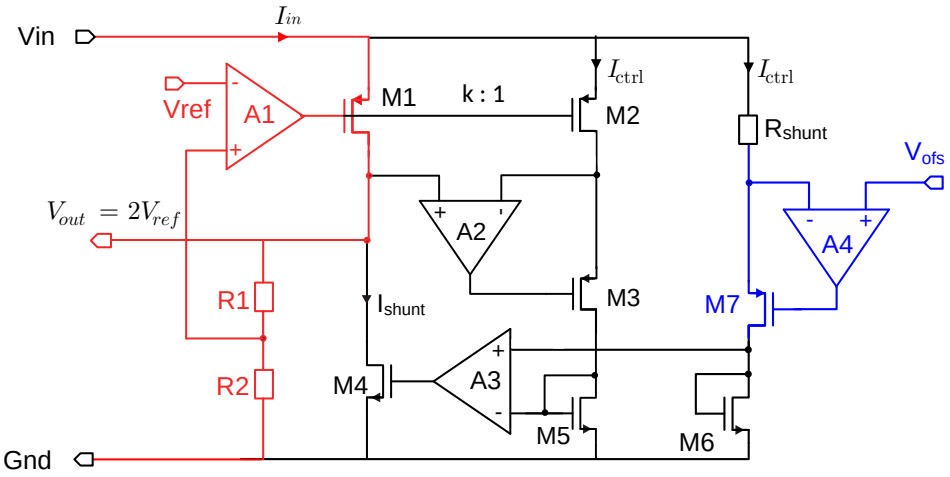


Figure 14: Simplified schematic RD53C SLDO regulator. The colors differentiate the LDO (red), shunt (black), and offset (blue) functions as discussed in the text.

from room temperature for bench testing and wafer probing, to the evaporative cooling base temperature (taken to be -40°C, that may be reached before power is applied. The generation of current and voltage references is intimately connected to the startup behavior.

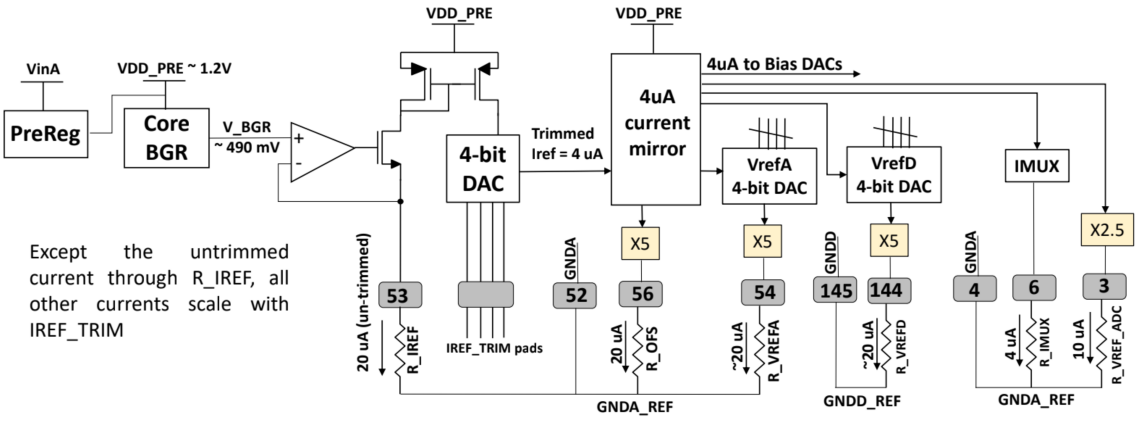


Figure 15: Generation of references and recommended connection to internal grounds. See Fig. 16 for further detail on R_OFS, shown here as a single resistor.

The RD53C reference scheme is shown in Fig. 15. RD53C does not use the SLDO output to power any reference circuit. A dedicated low current linear regulator (the *preregulator*) is used to power the main reference current generator. All other references are then derived from this unique main reference current. The preregulator is outside of and in parallel to the chip power delivered by the SLDO, but because it is low current it does not noticeably alter the behavior of Eq. 4.1 and Fig. 13. The preregulator includes its own dedicated bandgap voltage reference, which does not

need to be very precise, as the preregulator output does not need to be exactly 1.2 V, but merely between 1.1 V and 1.32 V. The preregulator is a low power device capable of a current of order 20 mA. This maximum current can be reached at very high shunt current I_s, well beyond normal operation values, and symptoms of a saturated pre-regulator can include increased jitter in the PLL leading to worsening of the output data eye diagram.

The Core Bandgap generates the main reference current, which can be adjusted with a 4-bit trim set by wire bond pads with internal pull-up resistors. This allows to compensate for process variations and equalize all chips to the design reference current value of 20 μA. Note that unless some of the pads are externally grounded the reference current will be at its maximum value. The generated main reference current also depends on the external resistor R_{Iref} (Table 30), which is external in order to avoid the temperature variation of internal devices. The two LDO reference voltages, V_{ref} analog and V_{ref} digital, are each generated by a known current (derived from the main reference) across a dedicated external resistor (Table 30). Each V_{ref} is independently adjustable by configuration to allow some fine adjustment of the chip internal operating voltage. V_{ref} adjustment does not change the V_{in} vs. I_{in} behavior of the chip, making sudden jumps due to configuration upset or operator error harmless for serial chain operation. The offset reference V_{ofs} is common to both SLDOs and is not adjustable by configuration, as sudden jumps in V_{ofs} would be problematic for serial chain operation. The generation is shown in Fig. 16.

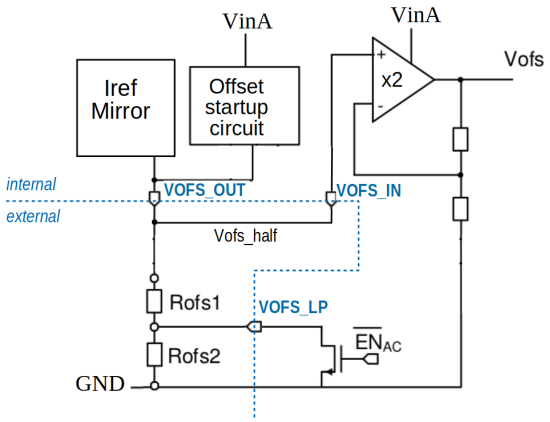


Figure 16: Offset voltage generation and startup. See text for description. The resistors in series Rofs1 and Rofs2 may sometimes be referred to as simply Rofs. The dashed blue line separates internal from external components and connections, while wire bonds pads are indicated by the pin symbol and blue labels. See Fig. 18 for optional connection of the VOFS_OUT and VOFS_IN wire bond pads on a module.

4.2.1 Offset Voltage Start-up

As the main current reference that all bias currents are derived from is powered from a dedicated linear regulator (the preregulator), circular dependencies requiring start-up circuits are generally absent from RD53C. However, the offset voltage (Vofs_half) does require a startup (Fig. 16), because a very low offset voltage would cause the shunt device (M4 of Fig. 14) to be fully on, and this would clamp the input voltage, V_{in}, to a low value even if a large current were supplied. This

is especially critical for low power mode (Sec. 4.3). The built-in start-up circuit shown in Fig. 17 boosts the offset voltage to follow V_{in} until V_{in} is high enough for the preregulator to work and all references to be at their correct values. The circuit injects a current into the offset voltage setting resistor until the preregulator reference voltage rises. The rise of the preregulator reference shuts off this startup circuit.

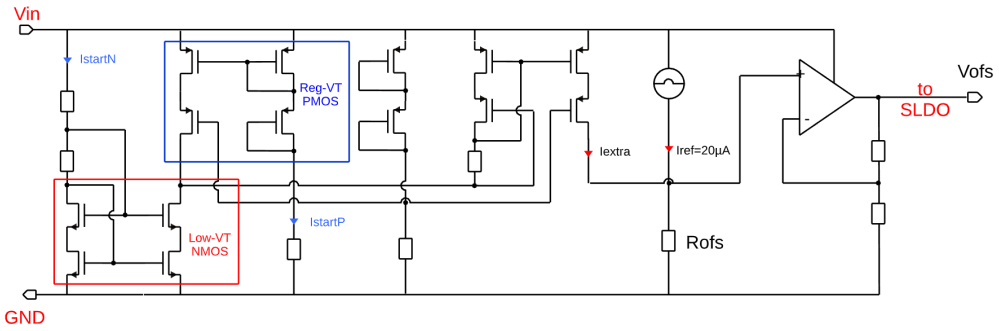


Figure 17: Offset voltage startup circuit. The Resistor Rofs is either Rofs1 in normal operating mode or Rofs1 + Rofs2 in low power mode (see Fig. 16).

4.3 Offset Voltage and Low Power Mode

The SLDO offset voltage plays critical roles. It is the most important voltage for regulators operating in parallel, because the total current in a given SLDO, I_{in}, is very sensitive to V_0 (and therefore to $V_{\rm ofs}$), as can be seen by the dI_{in}/dV_0 derivative to Eq. 4.1, which is -1/$R_{\rm eff}$. (This is also true for dI_{in}/dV_{in}, but V_{in} is by construction equal for all SLDOs wired in parallel.) It is very important to note that the actual offset voltage, V_0, in Fig. 13 is twice the generated Vofs_half voltage. This is because the actual offset voltage may need to be higher than the preregulator output voltage, and so is impossible to generate directly. A $\times 2$ buffer with $V_{\rm in}$ rail internally generates the true offset voltage from Vofs_half (Fig. 16). Only Vofs_half is accessible outside the chip and can be manipulated via the VOFS_OUT, VOFS_IN and VOFS_LP wire bond pads.

$V_{\rm ofs}$ may not necessarily be equal for different chips placed in parallel, and this can lead to current imbalance. While a small value of $R_{\rm eff}$ will make a single SLDO more efficient (lower voltage drop between V_{in} and V_{out}), it can make a multi-chip module less efficient by amplifying a small $V_{\rm ofs}$ chip-chip mismatch into a large current imbalance. Two solutions to this problem are possible in RD53C [2]. First, it is possible to trim the main current reference to produce a target $V_{\rm ofs}$ value, rather than to produce a target current value. This will result in a larger chip-chip variation of reference current, but since all internal biases are adjusted with dedicated DACs this is not a problem. In return for the larger variation of reference current there will be a smaller variation in offset voltage. The second solution is to tie together the $V_{\rm ofs}$ outputs of all chips in the same module via resistors (Fig. 18). For this purpose, in RD53C the $V_{\rm ofs}$ output of Fig 16 and the $V_{\rm ofs}$ input of Fig 14 are on separate wire bond pads.

[2] Assuming the $R_{\rm ofs}$ resistors cannot be practically trimmed individually

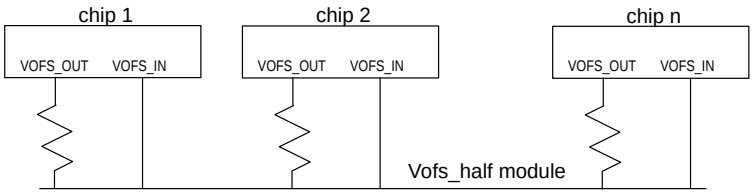


Figure 18: Common offset voltage wiring option for a multi-chip module. The VOFS_OUT and VOFS_IN wire bond pads can be seen in Fig. 16 and Table 22. For single chip operation VOFS_OUT should be simply looped back to VOFS_IN.

The common $V_{\rm ofs}$ wiring of Fig. 18 is robust against chip failure. Should one of the $V_{\rm ofs}$ outputs be grounded, the common $V_{\rm ofs}$ will be reduced. This will cause the working chips to draw more current for a given V_{in}, which is actually beneficial in the case the failing chip draws low current, as the working chips must now carry the extra current from the failing one.

Rather than a single external resistor to set $V_{\rm ofs}$, Fig 16 shows two resistors with a center tap switch- effectively a 1-bit variable resistor. The switch is internal in RD53C, while the resistors are external. This allows implementation of a low power serial chain mode. For normal serial chain operation the switch is conducting and the resistor value is just $R_{\rm ofs1}$. When the switch is off, the resistor becomes $R_{\rm ofs1} + R_{\rm ofs2}$, leading to a higher $V_{\rm ofs}$. A higher $V_{\rm ofs}$ means that a small current will develop a high enough value of V_{in} needed for the SLDO output to reach 1.2 V. This also requires the default configuration to be very low current, as is the case in RD53C ATLAS.

The switch is controlled by a dedicated A/C input as this mode is only intended for use during detector construction, when additional contacts can be made. An A/C signal on this special input will turn the switch off, and the absence of a signal (as will be the case for normal operation) will leave the switch conducting. Note that if this high $V_{\rm ofs}$ low power mode is never needed one can simply leave out $R_{\rm ofs2}$ and connect $R_{\rm ofs1}$ to ground, in which case it no longer even matters what the state of the switch is.

The amplitude of the A/C signal to activate low power mode should be 1.2 V peak-to-peak and should not exceed 1.32 V. A square wave with a rise time below 100 ns should have any frequency larger than 80 kHz, while a sine wave should have a frequency larger than 130 kHz. This assumes the A/C signal is coupled to the chip by a 100 nF external capacitor (Table 30). The rectification circuit that turns the A/C signal into an internal logic level is shown in Fig. 19. It consists of a 2 stage rectifier using low threshold NMOS transistors with applied forward-body biasing as rectification elements to achieve the minimum possible threshold voltage.

4.4 Under-shunt Current Protection

The variable internal shunt load L_s of Fig. 12 can act to keep the total current constant as long as the load current drawn by the chip, I_L, is less than the programmed total current I_{in}. But if due to an error or fault condition $I_L > I_{in}$, then there is nothing the variable load L_s can do to prevent the total current from exceeding I_{in}. An additional function is need to react to the condition $I_L > I_{in}$. The under-shunt circuit acts to prevent the $I_L > I_{in}$ condition. It is different from a classic current

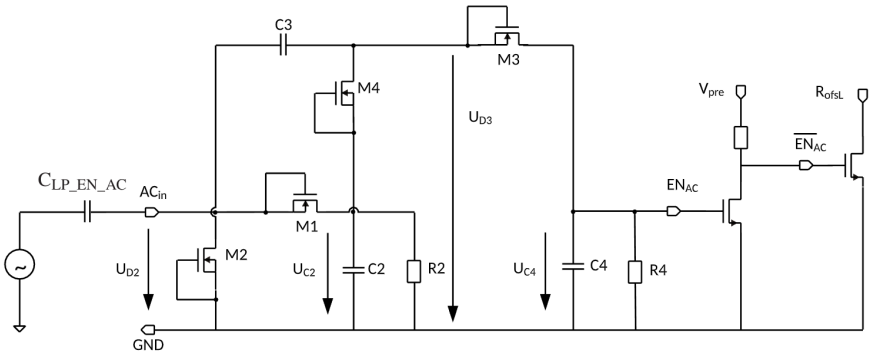


Figure 19: Rectification circuit for external A/C signal that enables low power mode.

limiting circuit, because the programmed value of I_{in} is not fixed in advance, but set by an external resistor. Thus it is not possible to have a hard-wired absolute current limit.

Turning around Eq. 4.1, $I_s = I_{in} - I_L$, where I_s is the internal shunt current in M4 of Fig. 14. The desired condition $I_L < I_{in}$ is equivalent to a non-zero shunt current, $I_s > 0$. Thus, the under-shunt protection compares a scaled replica of the M4 current to a threshold (which does not have to be precise), and if it goes below threshold (known as the under-shunt condition), it reduces V_{ref} (by reducing the current it is derived from). Lowering the voltage powering the load L will reduce the load current I_L. The circuit is shown in Fig. 20. However, the V_{ref} is not allowed to drop below 0.35 V, to avoid the possibility of a voltage greater than 1.32 V across M1 of Fig. 20, which could cause permanent damage to the device.

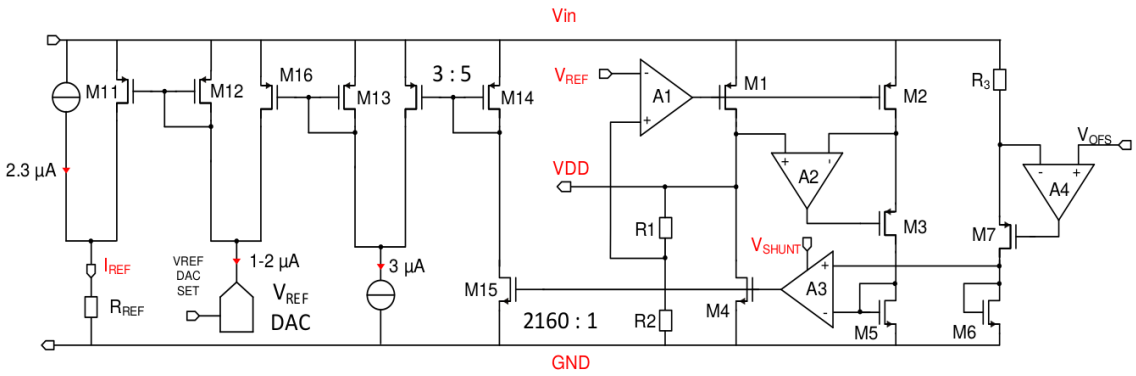


Figure 20: Under-shunt protection circuit.

The under-shunt protection is disabled by default and must be enabled in the global configuration. It can prevent internal shorts from being visible outside the chip, as long as their effective resistance is greater than 0.7 V/I_{in}. It can also prevent transient "shorts" (for example due to a simultaneous firing all comparators or a wrong configuration setting) from drawing more than the programmed I_{in}. Simulations of selected test cases show that the under-shunt protection generally mitigates both DC shorts and transients, but it can also lead to internal oscillation when the reduction of the load voltage removes the under-shunt condition, but then the condition returns when the load voltage recovers. These internal oscillations are not expected to be a problem for the system

outside the chip. Ultimately, the use or not of under-shunt protection will have to be informed by system tests.

4.5 Over-voltage Protection

In the SLDO design the shunt element M4 of Fig. 14 is placed after the pass device M1. The total current draw is limited by the pass device and additional current cannot be shunted by M4. Therefore, classic over-voltage protection (OVP) is implemented with a current clamp in parallel to the SLDO. Since the voltage being clamped is V_{in}, which is common to both SLDOs, there is only one single clamp for the whole chip. The circuit is shown in Fig. 21.

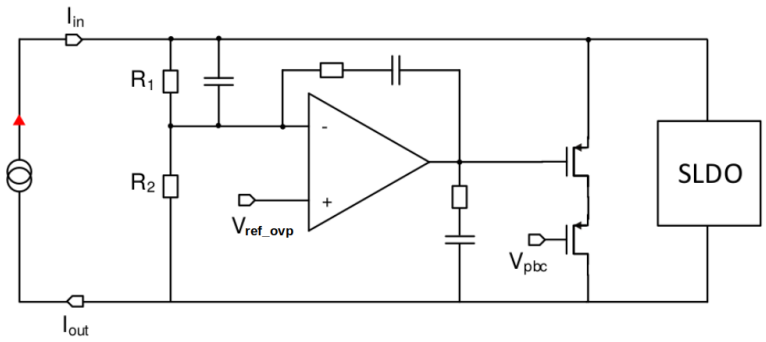


Figure 21: Over-voltage protection clamp.

The OVP must only become active if the input voltage is close to 2 V. The clamp threshold is $0.333 \times$Vref_ovp, where Vref_ovp is an internal reference obtained as a copy of the preregulator bandgap output voltage Vref_PRE and is expected to be around 0.6 V. This value can be overridden with the wire-bond pad VREF_OVP, without affecting Vref_PRE. OVP can be disabled by driving VREF_OVP to a high value (for example VDD_PRE). Note that if multiple chips in parallel go into OVP, there is no current balancing mechanism for this function, so the chip with the lowest effective OVP threshold will take most of the current.

5. Analog Front End

The ATLAS and CMS chips use different front ends (FE). However, they are treated the same way by the design framework and share many features. Much about the RD53C FE can be described generically, applying equally to ATLAS and CMS. The FE is a pure analog circuit: it contains no memory latches, flip-flops or counters. Static configuration values are provided by the digital core, which receives only the comparator output signal from the analog part. The design is a small-area, low-power, free-running amplifier and discriminator for negative input charge. All necessary biases are generated in the chip bottom as described in Sec. 5.1. The calibration charge injection circuit and operation are described in Sec. 6.

The FE circuits are laid out in analog islands of 4 FE's each, as was described in Sec. 1.

5.1 Front End Bias Generation and Distribution

The bias voltages for the analog front-ends are provided by a set of programmable 10-bit DACs placed in the Analog Chip Bottom, near the pad frame. The list of configuration registers of the bias DACs is available in Sec. 16.2.

The bias distribution to the pixel array is based on a 2-stage scheme, as shown in Fig. 22. The biases from the DACs are distributed in parallel to the DOUBLE_COLUMN_BIAS blocks placed at the bottom of the pixel matrix. Then, each DOUBLE_COLUMN_BIAS generates and distributes the bias and threshold voltages to two pixel columns. The chosen granularity allows a certain level of redundancy, so that a hard failure in one pixel will not affect the bias of the full pixel array.

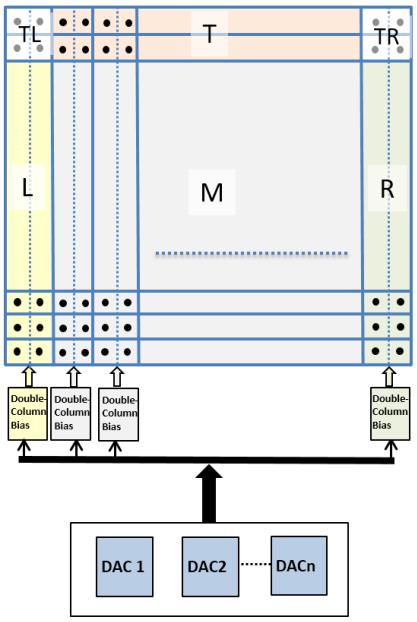


Figure 22: Bias distribution scheme.

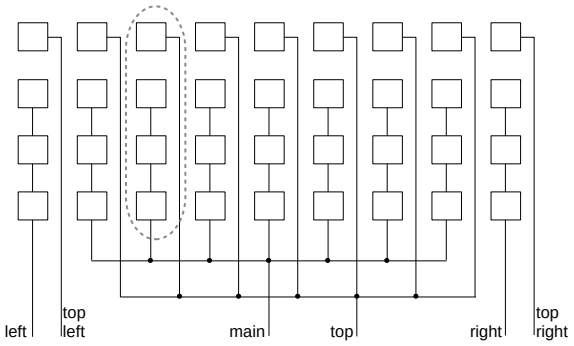


Figure 23: Clarification of input device bias scheme. Each square is a 2 by 2 pixel analog island. All columns are identical, with two bias lines (an arbitrary column is circled). Six DACs at the chip bottom control different lines as indicated. The distinction between center, sides, top, and corners is made by which DACs connect to which columns.

The distribution scheme takes also care to provide dedicated biases to the edge and top pixels, that will serve larger than normal sensor pixels to span the gap between adjacent chips in quad or dual chip modules. Edge/top pixels may need different bias to cope with greater capacitance and leakage current than the normal pixels. Simulations of the analog front-end showed that the only bias requiring different setting is the current of the preamplifier input transistor: a higher current allows both to align in time the response of the edge pixels and also partially recover the noise increase due to the greater capacitance. Therefore, only the input transistor bias can be adjusted differently for edge/top pixels, while all other biases are the same everywhere. The distribution of the input transistor bias is illustrated in Fig. 23. This distribution creates six groups of 4-pixel islands: Main (the interior if the chip), Left edge, Right edge, Top, Top Left corner, and Top Right corner. Each group has its own dedicated DAC for the input transistor bias, and these DACs can be set to the same or different values as needed by changing the value of the corresponding DAC_PREAMP_{M,L,R,TL,TR}_DIFF registers (ATLAS) or the DAC_PREAMP_{M,L,R,TL,TR}_LIN registers (CMS) (see Table 22).

In addition to the input transistor bias, the global thresholds setting is also modular for the left and right edge double-columns. Instead, the pixels of the top edge do not have an independent threshold. Therefore, the chip is equipped with three global threshold DACs: Left double column, Main array and Right double column. The global threshold is then set by setting a fixed value to the DAC_VTH2_DIFF register, and adjusting the value of the DAC_VTH1_{L,M,R}_DIFF registers (ATLAS) or adjusting the DAC_GDAC_{L,M,R}_LIN registers (CMS) as needed (see Table 22).

5.2 CMS Analog Front End

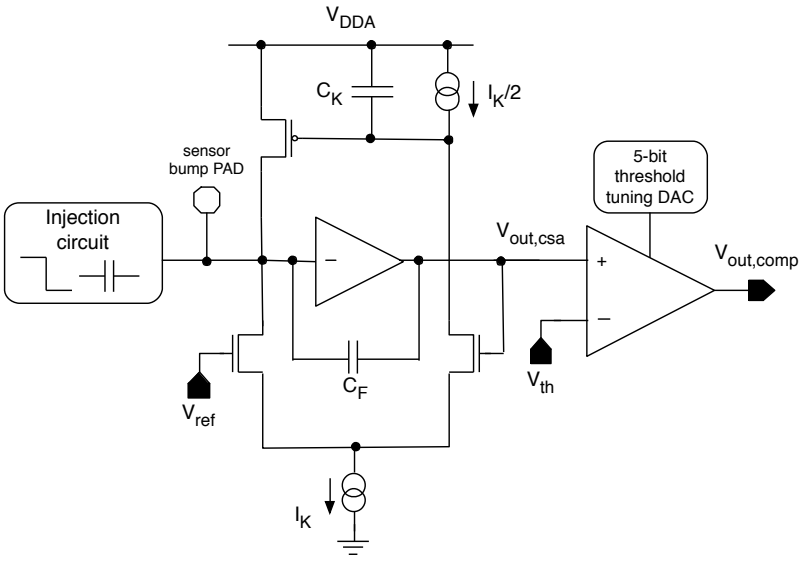


Figure 24: Schematic of the CMS analog front-end

The analog front-end (FE) for the CMS pixel readout chip is based on the RD53A Linear FE, with improved comparator and threshold-tuning DAC. The FE block diagram, with some transistor

level details, is shown in figure 24. The readout chain includes a charge sensitive amplifier (CSA) with Krummenacher feedback complying with the expected large radiation induced increase in the detector leakage current and providing a linear discharge of the feedback capacitor C_F. The designed charge sensitivity, set by C_F, is around 26 mV/ke$^-$. The signal from the CSA is fed to a low power comparator that, combined with a time-over-threshold counter, is exploited for time-to-digital conversion. Channel to channel dispersion of the threshold is addressed by means of a local circuit for threshold adjustment, based on a 5-bit, current-mode binary weighted DAC. The front-end chain has been optimized for a maximum input charge around 30 000 electrons and features an overall current consumption close to 5 μA.

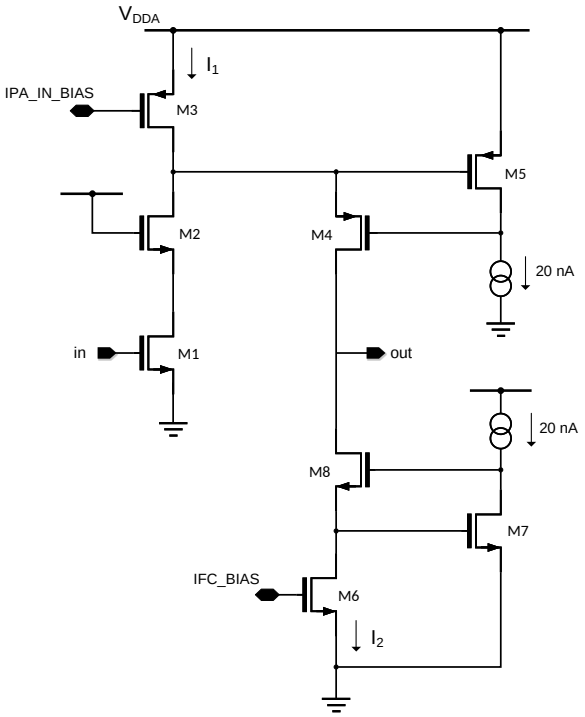


Figure 25: Charge sensitive amplifier forward gain stage.

The core element of the charge sensitive amplifier is the gain stage shown in figure 25. This is a folded cascode architecture including two local feedback networks, composed by the M4-M5 and M7-M8 pairs, boosting the signal resistance seen at the output node. With a current flowing in the input branch equal to 3 μA and a current in the cascode branch close to 200 nA, the CSA is responsible for most of the power consumption in the analog front-end. The DC gain and the -3dB cutoff frequency of the open loop response, as obtained from simulations, are 76 dB and 140 kHz, respectively.

The noise performance of the charge preamplifier is mainly determined by the contributions from the CSA input device and from the PMOS transistor part of the feedback network. The preamplifier gain stage is identical to the one integrated in the RD53A Linear front-end. Notice that

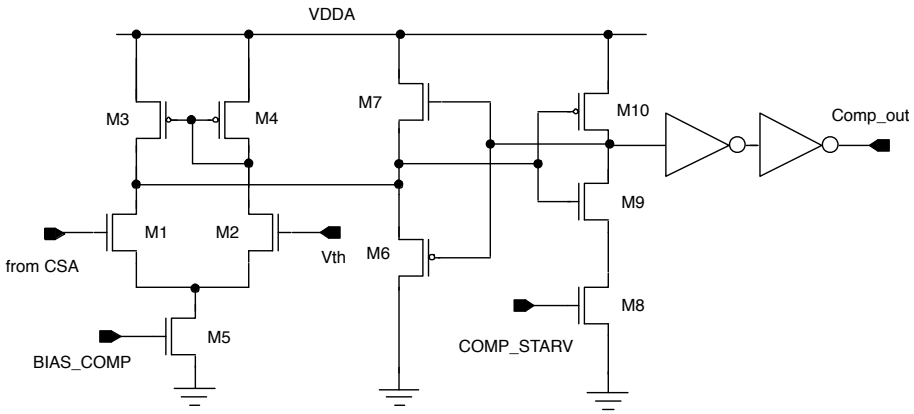


Figure 26: Threshold discriminator schematic diagram.

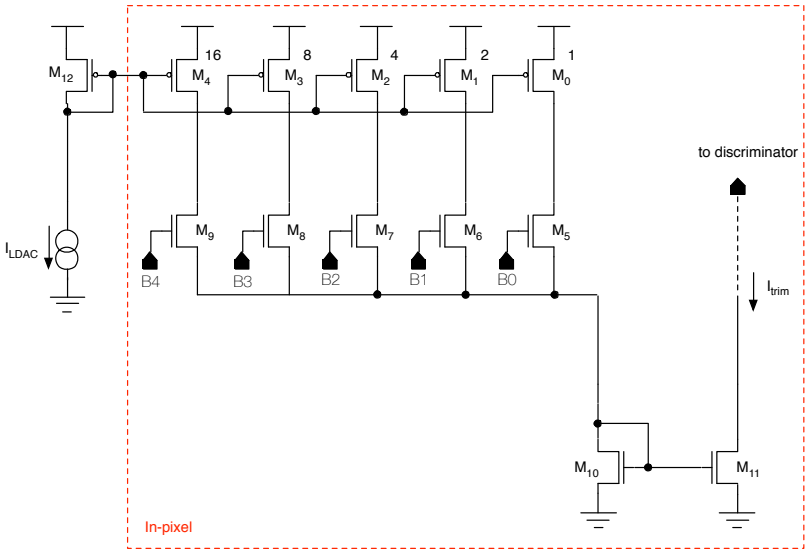


Figure 27: Threshold tuning DAC schematic diagram.

a gain selection bit was implemented in the RD53A version, whereas a single gain configuration is used in the CMS FE. Detailed simulation results are presented in [**?**].

The comparator schematic diagram is shown in figure 26. It includes a transconductance stage (M1-M5) whose output current is fed to the input of a transimpedance (TIA) amplifier providing a low impedance path for fast switching. A couple of inverters is used after the TIA in order to consolidate the logic levels. With respect to RD53A, the main modifications to the comparator are in its transimpedance stage. In particular, the TIA feedback network (transistors M6 and M7) has been modified with the aim of improving the time-walk performance of the front-end. A starving mechanism (transistor M8) has been added in the gain stage (M9 and M10) to limit the power consumption of the TIA. The trimming DAC schematic diagram is shown in figure 27. This is a 5-

Table 4: Linear AFE settings

	Rec	Std Range
FC_BIAS_LIN	20	10-70
Vthreshold_LIN	-	REF_KRUM_LIN-900
COMP_LIN	110	70-250
COMP_STAR_LIN	110	50-300
LDAC_LIN	110	80-COMP_LIN×2
KRUM_CURR_LIN	70	5-200
PA_IN_BIAS_LIN	300	100-700
REF_KRUM_LIN	360	300-450

bit, current-mode, binary weighted DAC featuring regular current mirrors. The DAC output node, sinking a current I_{trim}, is connected to the output of the transimpedance stage of the comparator.

5.2.1 Front-end registers

Table 4 reports a list of the DAC settings for the CMS AFE, showing the recommended (Rec) values and the operating range (Std range). The recommended values are used to operate the front-end with a total analog current of 5 μA, with a ToT close to 130 ns for an input charge of 6000 electrons. Some simulation results are given in this section and more detailed results can be found in [**?**]. The DAC settings can be divided into two categories, in particular:

- Main settings:
  - Vthreshold_LIN
  - LDAC_LIN
  - KRUM_CURR_LIN
- Secondary settings
  - FC_BIAS_LIN
  - COMP_LIN
  - COMP_STAR_LIN
  - PA_IN_BIAS_LIN
  - REF_KRUM_LIN

The main DAC settings can be modified to achieve different threshold and ToT operation, whereas secondary settings should be in principle kept at their default value.

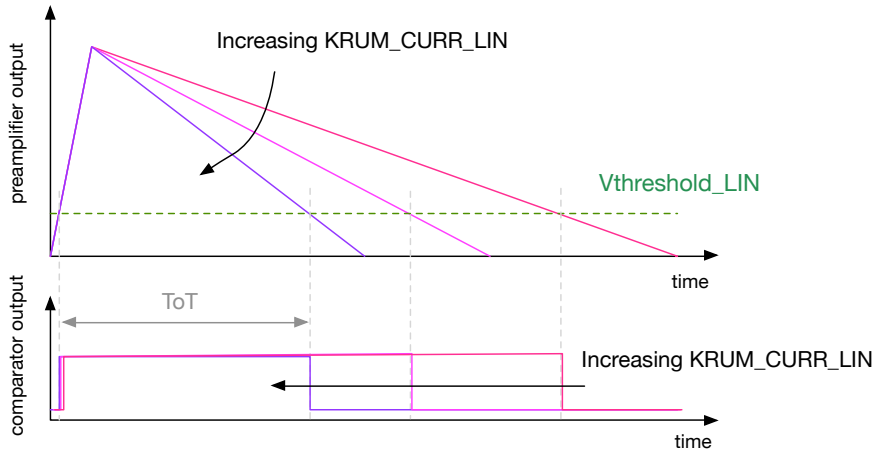


Figure 28: Effects of KRUM_CURR_LIN on preamplifier output waveform and ToT.

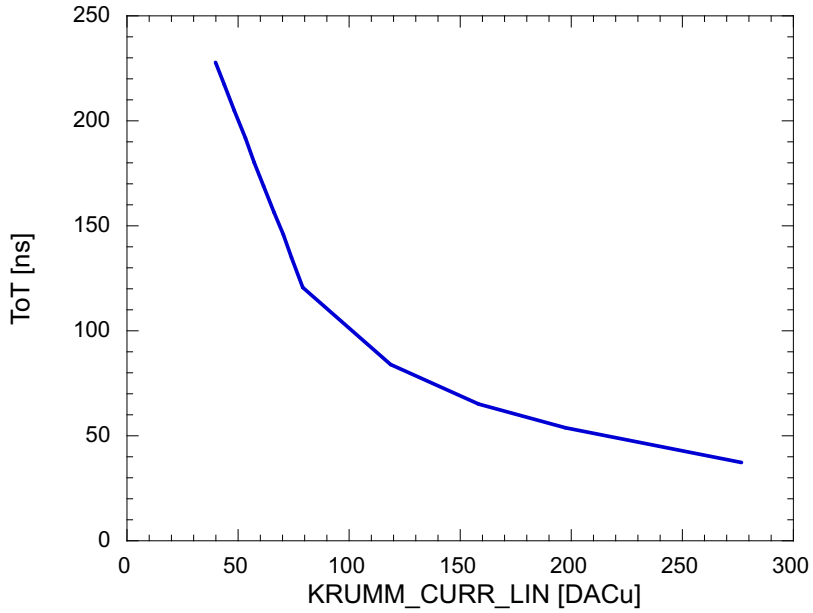


Figure 29: ToT as a function of KRUM_CURR_LIN for a 1000 electrons threshold.

5.2.2 Krummenacher current: KRUM_CURR_LIN

KRUM_CURR_LIN sets the current in the Krummenacher feedback, used to linearly discharge the preamplifier feedback capacitance. Increasing KRUM_CURR_LIN results in a faster preamplifier return to baseline and a reduced Time Over Threshold (ToT), as schematically shown in Fig. 28. The recommended value, 70, should results in a ToT close to 133 ns for an input charge of 6000 electrons. The simulated relationship between KRUM_CURR_LIN and the ToT is shown in Fig. 29, for a threshold set around 1000 electrons.

5.2.3 Global threshold: Vthreshold_LIN

Vthreshold_LIN sets the global threshold of the Linear AFE, corresponding to the DC thresh-

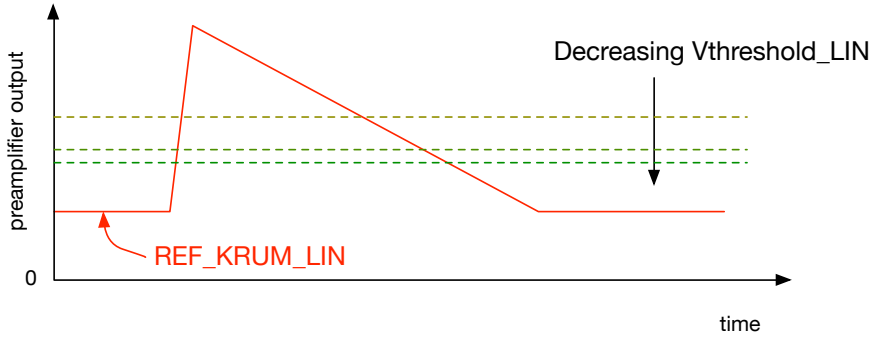


Figure 30: Effects of Vthreshold_LIN on the global threshold.

old voltage applied to the discriminator input. Increasing Vthreshold_LIN results in an increased global threshold, as conceptually shown in Fig. 30. It is worth noticing that the effective threshold is the difference between Vthreshold_LIN and REF_KRUM_LIN which, in turn, sets the preamplifier output DC baseline (increasing REF_KRUM_LIN results in a higher baseline). Vthreshold_LIN=400 should provide a threshold (after tuning) close to 1000 electrons (as REF_KRUM_LIN is kept to 360).

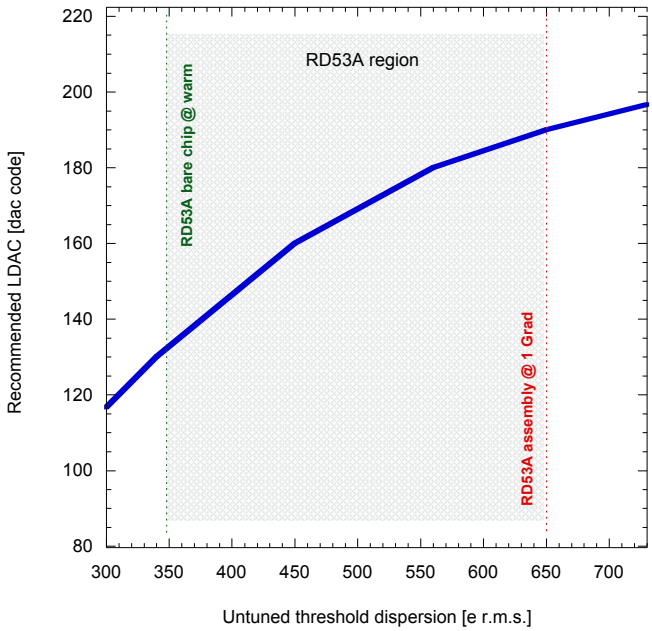


Figure 31: Recommended LDAC_LIN values as a function of the un-tuned threshold dispersion.

5.2.4 Threshold trimming: LDAC_LIN

LDAC_LIN sets the LSB of the in-pixel threshold trimming DAC and, in turn, its output dynamic range. In particular, LDAC_LIN controls the current I_{LDAC} shown in the schematic of Fig. 27. Increasing LDAC_LIN results in increased LSB and output range. An increase in the un-tuned

threshold dispersion is expected for the Linear AFE after irradiation. In that case it could be required to operate the front-end with an increased LDAC_LIN in order to compensate for the larger dispersion.

Recommended values of LDAC_LIN are shown in Fig. 31 as a function of the un-tuned threshold dispersion. Typical values of the untuned threshold dispersion before irradiation are in the range 350-550 electrons (as obtained from RD53A and prototype measurements, depending on sensors geometry, temperature, ...). As shown in the figure, the untuned threshold dispersion can get up to 650 electrons after irradiation at 1 Grad (from RD53A measurements).

5.2.5 Secondary settings

The DAC settings listed in this section are mainly related to preamplifier and comparator bias and should be, in principle, kept at their default value. A brief overview of these settings is given in the following.

- FC_BIAS_LIN: sets the current in the preamplifier folded cascode branch.
- COMP_LIN: sets the bias current in the threshold discriminator input (transconductance) stage. This parameter can be slightly increased to improve speed.
- COMP_STAR_LIN: sets the maximum current flowing in the comparator output branch (controlled by the starving transistor M8 in Fig. 26).
- PA_IN_BIAS_LIN: sets the current in the preamplifier input branch. This current represents the main contribution to the Linear AFE current consumption. This parameter can be slightly decreased to reduce power, at the cost of an increased noise.
- REF_KRUM_LIN: as explained in section 2.2, this parameter sets the preamplifier output DC baseline ($V_{out,csa}$ in Fig. 24).

6. Calibration Injection

The calibration injection circuit can internally inject signals into any combination of pixels without the need for a sensor or radiation. There are two types of calibration injection: digital and analog. The same command (Sec. 6.2) is used for both, and which one is active is selected by configuration in register CalibrationConfig (Table 22). Digital injection bypasses the front end and inputs a digital pulse to the hit processing logic as shown in Fig. 41. It is therefore relatively simple: the digital pulse generated by the Cal command is directly what the hit processing uses and is fully deterministic (no noise). It is useful to test proper functioning of the readout chain, as a timing reference for each pixel's FE analog delay, etc. The rest of this section is concerned with analog injection.

The calibration injection circuit uses two distributed DC voltages plus in-pixel switches to chop them and generate steps fed to an injection capacitor. Having two voltages allows a precise differential voltage that will be independent of local ground drops in the chip, as well as two consecutive injections into the same pixel. The injection circuit is implemented in every pixel and its topology is shown in Fig. 32. The control signals, S0 and S1, are generated in the digital domain as explained in Sec. 6.1. They can be phase shifted relative to the bunch crossing clock with a fine delay, which is global for the whole chip. The enable bit (EN) is programmable for each pixel and injection takes place only for enabled pixels. Charge is injected when either S0 or S1 switch from low to high. Analog injection must therefore be primed by setting at least one control signal low, prior to being able to inject. This priming is not automatic, so that the user is able to control the amount of settling time allowed prior to injection. The CAL command is used for both functions: prime and inject (see Sec. 6.2). The value of the injection capacitor can deviate from nominal due to process variations, so a dedicated circuit is provided to measure a replica capacitor array in each chip during wafer probing (Sec 13.8).

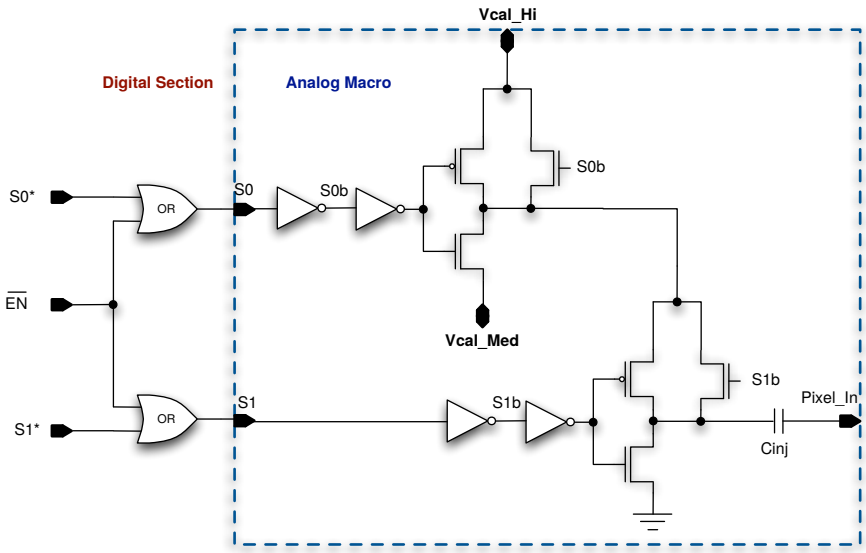


Figure 32: Calibration injection circuit in each pixel. The injection capacitor nominal value is 8.02 fF.

The injection circuit resembles two CMOS inverters, and just as in a common inverter, there will be a switching transient when a control signal switches from low to high, but simulations show these transients to have a negligible impact on the distributed Vcal_Hi and Vcal_Med voltages. Note that at the top of each inverter the injection circuit adds an NMOS transistor in parallel with the PMOS, which switches first (before the PMOS) when injecting. This allows the switches to operate for any choice of voltages Vcal_Hi>Vcal_Med>GND, but since the top NMOS switches first, it does not contribute transients during injection. During priming, on the other hand, the top NMOS switches first, while the bottom NMOS is still conducting, resulting in a short circuit lasting one inverter delay. This will cause a transient on the Vcal_Hi and Vcal_Med voltages, and the user must therefore allow some settling time between priming and injection. In addition to this transient, priming injects a positive polarity pulse into each enabled front end, so one must allow for the front end to settle in any case. The use of two voltages means that the charge injected by S0 is given by a differential voltage and not affected by local ground potential differences. Keeping S1=0 and only toggling S0 will result in single pulse differential injection. Conversely, the two-voltage injection circuit also allows injection of two successive pulses without priming in between, and with arbitrary delay between these pulses (one pulse differential and another referenced to ground). An additional control feature exploits the use of two voltages to inject a different amount of charge simultaneously in neighboring pixels, by changing the meaning of S0 and S1 in different pixels (see Sec. 6.1).

Finally, since the voltage distribution lines have finite impedance, injecting into too many pixels simultaneously will cause the voltages to droop, introducing a nonlinearity in the injected charge vs. number of pixels injected. Simulations show this nonlinearity to be less than 1% for simultaneous injection into 3 full rows of pixels and less than 2% for 4 rows, but the exact value of this nonlinearity should be measured in actual chips by measuring threshold vs. number (and pattern) of injected pixels.

6.1 Generation of S0 and S1 signals

The signals S0 and S1 of Sec.6 exist locally in each pixel but are derived from different internal signals produced by the command decoder and distributed to the array. This two-step scheme is necessary in order to implement the above injection options of either consecutive different pulses into the same pixel or parallel different pulses into neighboring pixels. It also avoids having to distribute two switching signals with precise timing, instead of just one, saving power and area. Since the calibration input is used to study and calibrate timing, it must occur simultaneously in all enabled pixels, just as is the case for the bunch crossing clock (here, simultaneously means within a 2 ns window). Two control signals are distributed: CAL_edge and CAL_aux. As the name implies, CAL_edge needs to be simultaneous in all pixels, while CAL_aux does not. CAL_edge has a fine phase adjustment relative to the beam crossing clock, which is called CAL_delay. In fact CAL_edge looks like a traditional injection pulse, with user controlled leading edge time and duration. In *uniform* injection mode (which allows injecting two pulses close in time into all selected pixels), S0 and S1 are derived from CAL_edge and CAL_aux identically for all pixels:

$$\mathrm{S0} = \mathrm{CAL\_edge}\ \ \mathit{OR}\ \ \mathrm{CAL\_aux} \tag{6.1}$$

$$\mathrm{S1} = \overline{\mathrm{CAL\_edge}}\ \ \mathit{AND}\ \ \mathrm{CAL\_aux} \tag{6.2}$$

The rising edge of CAL_edge throws the S0 switch, while the falling edge throws the S1 switch. The CAL_aux starts low and then goes high after CAL_edge, but not with precise timing. In uniform mode the injection switches can only be thrown in that order. Either only use S0 for single pulse, differential voltage injection, of use S0, then S1, for double pulse injection.

In order to allow injection of different size pulses simultaneously into adjacent pixels, there is an *alternating* analog injection mode that can be selected instead of the default uniform mode. In this mode the S0 and S1 signals are derived as above only for *even* pixels, but swapped for *odd* pixels:

$$\mathrm{S1}_{odd} = \mathrm{CAL\_edge} \;\; OR \;\; \mathrm{CAL\_aux} \tag{6.3}$$

$$\mathrm{S0}_{odd} = \overline{\mathrm{CAL\_edge}} \;\; AND \;\; \mathrm{CAL\_aux} \tag{6.4}$$

where an even (odd) pixel is one for which the sum of row + column is an even (odd) number. Thus, for example, in single injection mode the CAL_edge rising edge throws S0 for even pixels, but S1 for odd pixels. The S0 and S1 assignment options are independent of the cal enable bit in each pixel. The Analog Mode bit of the injection configuration controls whether injection is uniform (mode=0) or alternating (mode=1).

6.2 Cal Command

The Cal command controls the generation of the two internal signals CAL_edge and CAL_aux. For digital injection only the CAL_edge signal is relevant. The CAL_edge signal to be generated is specified by the first 14 data bits of the Cal command, while the CAL_aux signal is specified by last 6 data bits. The detailed bit assignment of the command payload (four 5-bit fields) is shown in Fig. 33.

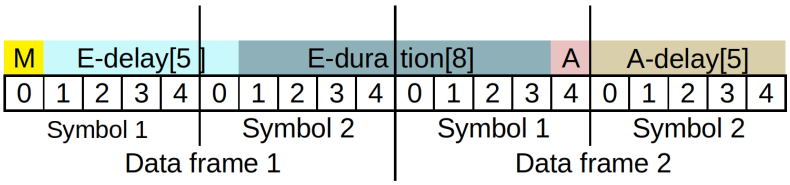


Figure 33: Bit assignment of the Cal command payload. Two data frames totaling 20 bits. M= mode bit, E= CAL_edge parameters, A= CAL_aux parameters (value and delay). All delays and duration are in units of 160 MHz clock cycles.

The CAL mode bit (M) selects between two behaviors for the CAL_edge signal: a single step (mode=0) or a pulse (mode=1). The step is always from low to high, so an edge is only produced if the prior state of CAL_edge was low; if it was high it remains high. Thus, for typical injection it is necessary to first arm the system to ensure CAL_edge is low, as shown in Fig. 34. The standard injection sequence is then shown in Fig. 35. In this case voltage at the injection capacitor of the selected pixels is switched from Vcal_Hi and Vcal_Med, effectively providing a differential injection voltage that will be insensitive to power and ground local voltage variations across the matrix. The delay value controls the “coarse” delay from the Cal command to the injection (in cycles of the 160 MHz clock), and a global fine delay is added on top of that. This fine delay is

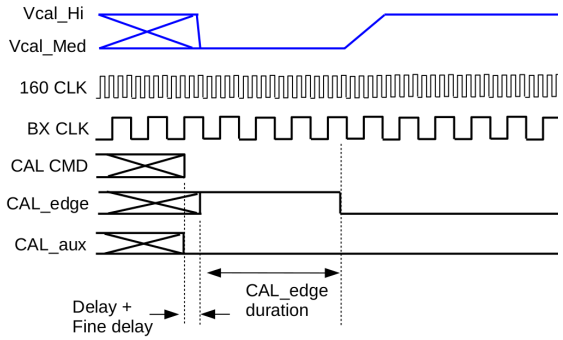


Figure 34: Timing diagram illustrating the arming of calibration injection, to set both CAL_edge and CAL_aux to the correct levels without knowing their prior state. The Cal command parameters are as follows: M=1, delay=1, duration=15, A=0, Adelay=0 (the exact delay and duration values are not important).

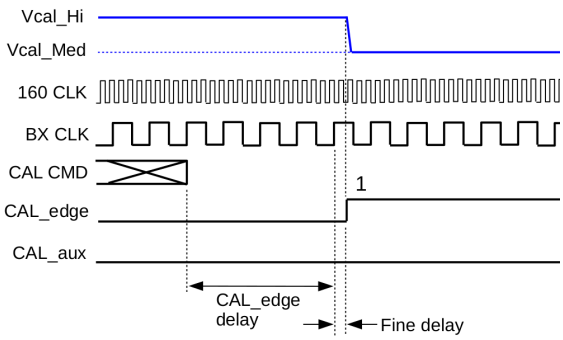


Figure 35: Timing diagram illustrating standard calibration injection. Only the CAL_edge signal is active. The Cal command parameters are as follows: M=0, delay=16, duration=1, A=0, Adelay=0 (the exact delay and duration values are not important).

pre-programmed in a global register and is in units of 1.28 GHz clock cycles. This allows precision scanning of the pixel timing. The duration serves no purpose in edge mode and the CAL_aux signal is unchanged. But note that CAL_aux must be low the whole time in order for the rising CAL_edge to switch from Vcal_Hi and Vcal_Med. If the CAL_aux signal were instead high, the switching would be from ground to CAL_edge. This would inject positive polarity charge, which the Front End is not designed for, but may be of interest for special tests.

In pulse mode (M=1) the CAL_edge signal will be set to high after the coarse plus fine delay, just as it happened for edge mode, but then it will be set to low after the duration value elapses. Note that if duration is set to zero then CAL_edge will simply go low after the delay (a duration zero pulse with final state low). Thus, pulse mode with duration zero is the complement of step mode: the former brings CAL_edge low while the latter brings CAL_edge high. Step mode is used to produce two consecutive injections. Starting from the armed state of Fig. 34, the rising CAL_edge will inject from Vcal_Hi and Vcal_Med as usual, but then, before the falling CAL_edge at the end of injection the CAL_aux signal is set high, which then causes the falling CAL_edge to switch from Vcal_Med to ground. This sequence is shown in Fig. 36. The CAL_aux signal will be set to the level indicated in the command (A) after the given delay value. The fine scale for the delay allows changing the CAL_aux value in the middle of a bunch crossing cycle, as is needed to inject charge in two consecutive crossings. Since at the end of this sequence CAL_edge is low, it's no longer necessary to repeat the Fig. 34 arming sequence. However, the CAL_aux signal has to be returned low, which can be done with the sequence in Fig. 37.

Executing a double injection with a single Cal command in pulse mode is limited the injections being closer together in time than the maximum CAL_edge duration. To perform a double injection separated by longer times, two separate Cal commands can be used. The first would be a standard injection (Fig. 35), while the second command would accomplish the Vcal_Med to ground injection as shown in Fig. 38. Note that this is not the same as two consecutive standard

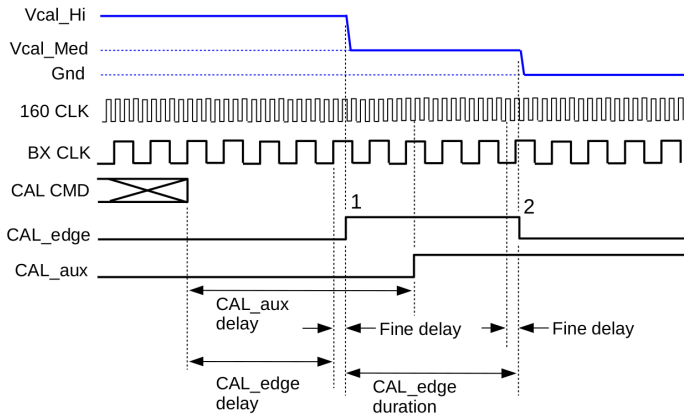


Figure 36: Timing diagram illustrating double calibration injection using the pulse mode of the Cal command. The Cal command parameters are as follows: M=1, delay=16, duration=19, A=1, Adelay=25 (the exact values are representative. The Cal_AUX transition must be between the CAL_edge rising and falling edges).

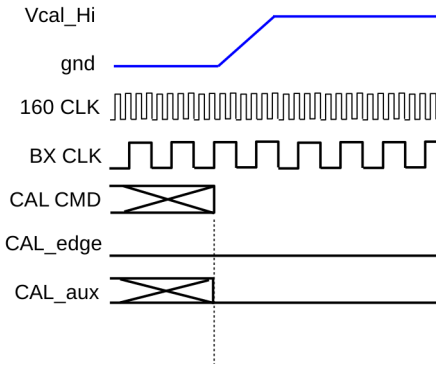


Figure 37: Timing diagram illustrating the re-arming of the CAL_aux signal after double injection. The Cal command parameters are as follows: M=1, delay=0, duration=0, A=0, Adelay=0.

injection sequences, because in between standard injections one must re-arm, which produces a positive polarity injection at the time of re-arming. In contrast, in the double injection there is no activity between the two injections. This can be important when studying threshold stability vs. time, for example. One can achieve the same thing using standard injection, but it requires more commands and, therefore, more time between injections.

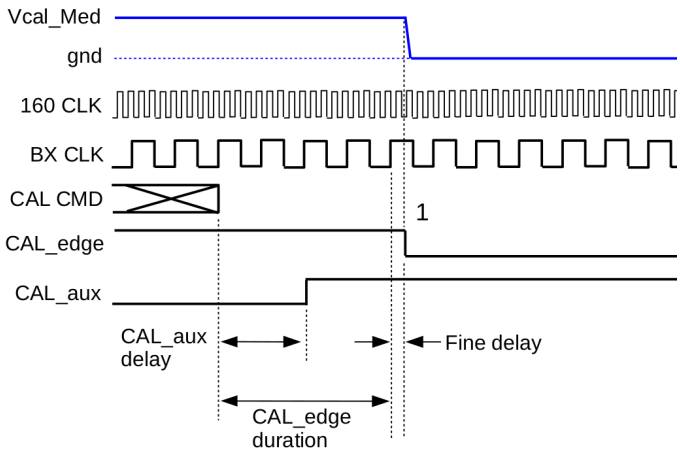


Figure 38: Timing diagram for an arbitrarily delayed second injection in a double injection sequence. The Cal command parameters are as follows: M=1, delay=0, duration=16, A=1, Adelay=8 (the exact values are representative. The Cal_AUX transition must be before the CAL_edge falling edge).

6.3 Injection Voltages

The two injection voltages Vcal_Hi and Vcal_Med are generated by two 12-bit DACs in the chip bottom as shown in Fig. 39. Using the SEL_GRANULARITY configuration bit, the circuit can operate either with high dynamic range or with fine step sizes. In the former configuration, the DACs voltage reference is the same as the ADC voltage reference Vref_ADC (Sec. 12.2), while in the latter case it is Vref_ADC/2. As these voltages are relatively high impedance, injection into many pixels at once will introduce a systematic bias, as the current pulse from the combined effect of all injected pixels will cause a voltage shift of the injection voltages. This effect is small and can be ignored for many applications, but should be considered for precision studies. An additional consideration for any precision studies is that the absolute scale of calibration injection depends on Vref_ADC and on the injection capacitor value, C_{inj}. Therefore, when converting to units of injected charge (typically in electrons), the conversion is only as good as the knowledge of Vref_ADC and C_{inj}. This conversion is typically done automatically within the readout systems. The value of Vref_ADC (which depends on the trimmable Iref main current) should be checked and C_{inj} should be measured (Sec. 13.8).

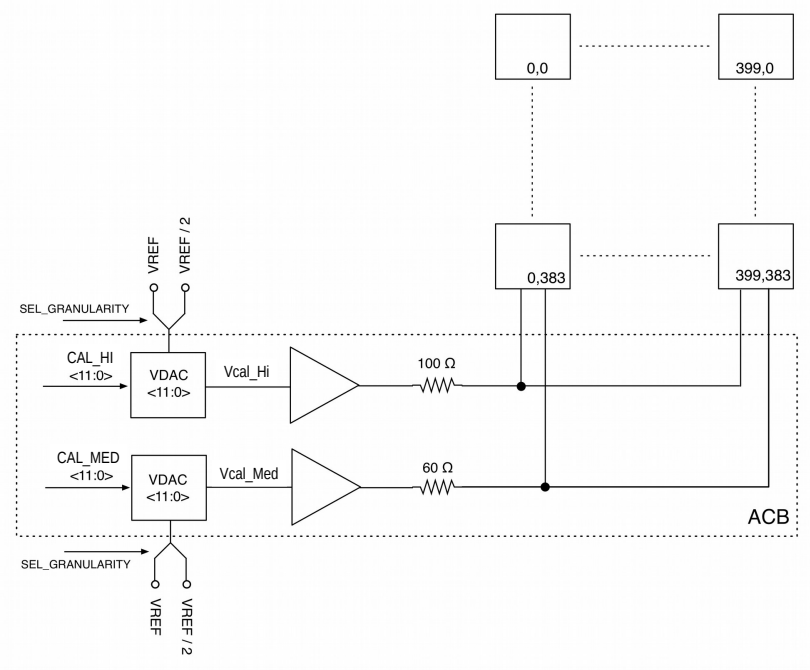


Figure 39: Generation of the injection voltages Vcal_Hi and Vcal_Med.

7. Digital Core

The digital core implements all the functionality of digitization, time stamping, storage, trigger retrieval, configuration memory, and injection control for the 64 pixels in the core. All cores are identical and the layout is stepped and repeated to from the matrix. The address of each core is generated by a combinatorial subtracter that subtracts 1 to the address of the previous core, starting from a hard-wired seed address of 41 at the bottom of the matrix (thus the top-most core has address 0). The distribution of the bunch crossing clock and calibration pulse (CAL_edge) along the column is not done with a global clock tree, but they are time aligned though delays that depend on the core address, compensating for the propagation delay accumulated along the column.

Within the core, the pixel hits are processed in pixel regions (Sec. 7.1), each made of 4 pixels (i.e. in total there are 16 pixel regions in a digital core). Even though each analog island also has 4 pixels, the region pixels are organized in 4×1 rows, a region receives the hits from two pixels in one analog island and two in another. The 4 pixels in a region share timing information, while each individual pixel has its own hit processing (Sec. 7.2) along with dedicated ToT counter and memory (Sec. 7.3). The pixel region aspect ratio is chosen to minimize the size of its timestamp and ToT memories for the areas of the detector with the highest hit rates (i.e. the end of the innermost barrel, where a particle normally hits an elongated cluster of pixels). Additionally, the region manages clock gating, which reduces digital power consumption even at the maximum hit rate. This is because even at the maximum hit rate not all pixel regions are processing hits at any given point in time. At maximum hit rate (3 GHz/cm^2), digital power is approximately half of what it would be without clock gating.

7.1 4-Pixel Region

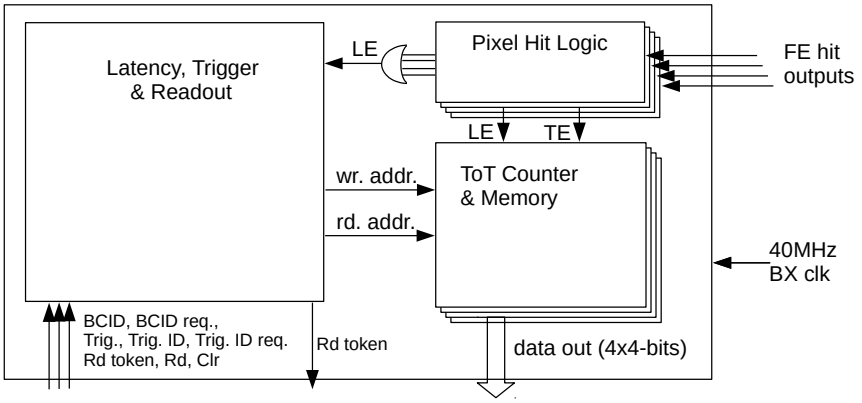


Figure 40: 4-Pixel region block diagram. LE is leading edge, TE is trailing edge, BCID is bunch crossing counter value. See text for details.

The pixel region logic contains three main blocks as indicated in Fig. 40: the Hit Logic (four instances), the ToT counter and storage (four instances), and one Latency, Trigger and Readout

(LTR) block. Sharing the LTR block among four pixels leads to a more compact layout than if each pixel was independent.

The Hit Logic (Sec. 7.2) determines if an hit is present in a given bunch crossing and drives the ToT counter (Sec. 7.3). The LTR block (Sec. 7.4) is the brain of the region. It keeps track of the timing of all hits, decides which ToT memories to use, and manages triggering, reading and clearing of ToT data.

7.2 Pixel Hit Logic

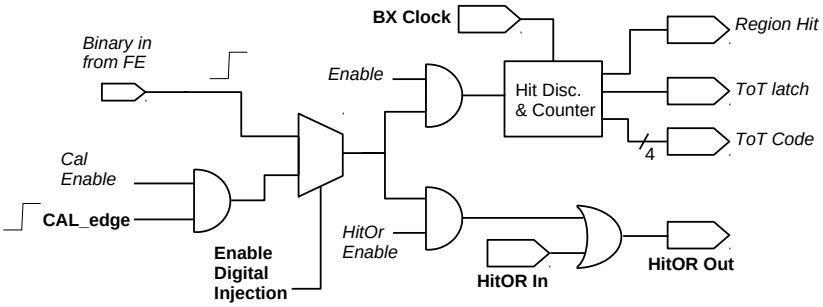


Figure 41: Single pixel hit digital processing path. The bold text indicates global signals, while the italic text indicates local pixel signals.

The hit output of the analog front end is processed to produce all the needed digital values and signals as shown schematically in Fig. 41. There are two parallel paths with independent enable bits: the DAQ path leading to hits being collected and encoded into the chip data output (Sec. 10, and the HitOr path, which feeds the wired OR core column lines as described in Sec. 13.4. The hit source can be selected to be the analog front end, or digital hit calibration injection, which checks the full digital functionality bypassing the analog front end. The Cal Enable, (data output) Enable, and HitOR Enable bits are independently set for every pixel (Sec. 8.8), while Enable Digital Injection is globally controlled by the CalibrationConfig register (Table 22) The Hit Disc. & Counter block contains logic to detect and synchronize hits (Fig. 42), as well as a ToT counter dedicated to this pixel (Sec. 7.3).

Fig. 42 shows the schematic of the "asynchronous option" logic to process the hit output of the analog front end, while Fig. 43 shows the corresponding waveforms. This option will detect all FE output pulses even if they are very short duration. An alternate "synchronous option" is also available. If selected, this option will fist synchronize the FE output with the 40 MHz clock before further processing. Now only output pulses that are high during a clock transition will be registered as hits. Thus all pulses wider than one clock cycle will be registered, while pulses narrower than one clock cycle will sometimes be registered and sometimes not, depending on their phase relationship to the clock.

7.3 ToT counter and storage

ToT is counted independently for each pixel using the 40 MHz clock. An 80 MHz effective resolution is provided by capturing the clock phase with the asynchronous hit. In this case, the clock

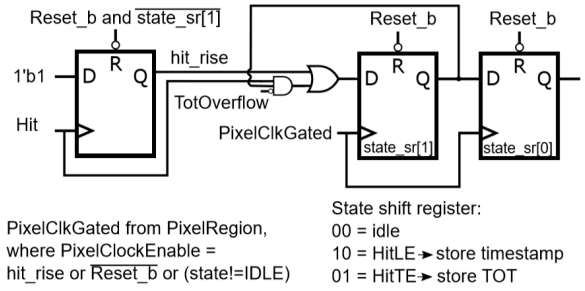


Figure 42: Schematic of the logic to process the asynchronous hit output of the analog FE. The reset_b signal globally provided during initialization. The state shift register controls clock gating, timestamping, ToT counting and storage.

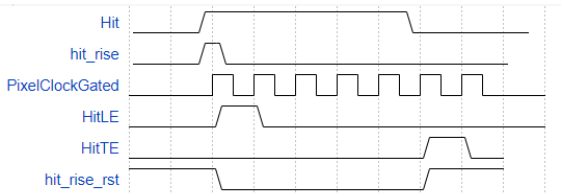


Figure 43: Waveforms showing the processing of a hit output from the analog FE. HitLE stores the timestamp in the pixel region, HitTE stores the ToT in the pixel ToT memory.

falling phase is latched, based on the assumption that hits are time-aligned on the rising-edge. 80 MHz ToT counting can be enabled through global configuration (the default is 40 MHz). The ToT counter schematic is shown in Fig. 44. A 6-bit counter is used (including the above mentioned clock phase capture bit), but only 4 bits are stored and read out, as described below.

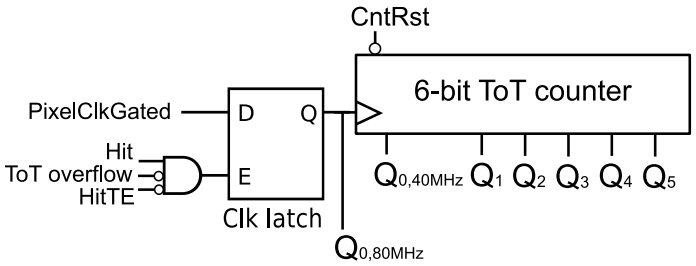


Figure 44: ToT counter schematic. At the end of the count, the Clk latch captures the clock falling phase, which is used as the LSB ($Q_{0,80\,\mathrm{MHz}}$) of the count, achieving 80 MHz ToT counting. The input signals reported on the schematic are the same shown in the pixel control logic in Fig. 42.

In addition to 80 MHz resolution, a dual slope ToT mapping is supported and can be enabled by global configuration. In this case, the 6 bits of the counter are compressed to 4 bits. If the feature is disabled, either the 4 LSBs (80 MHz mode) or the middle 4 bits (40 MHz mode) are stored. The logic implementing the dual slope mapping is shown in Fig. 45.

The meaning of each 4-bit ToT code in terms of true ToT value is shown in Table 5 for the two possible speeds or dual slope compression. Note that in the default mode (40 MHz and no compression) the output ToT code is the true ToT bin low edge.

In any mode, the ToT code that is read out goes from 0 to 14. Code 15 is reserved in the pixel region to identify non-hit pixels. Also, if the ToT counter reaches maximum while the pixel comparator output is still high, the counting concludes and the maximum ToT (14) is recorded. The pixel region clock is gated off whenever there is no ToT counting taking place.

The ToT storage has 8 locations per pixel, 4 bits each. The value of the ToT counter is stored

Output 4-bit code	True ToT bin (low edge) [BX]			
	40 MHz speed		80 MHz speed	
	4-bit (DEF)	6-to-4 bit	4-bit	6-to-4 bit
0	0	0	0	0
1	1	1	0.5	0.5
2	2	2	1	1
3	3	3	1.5	1.5
4	4	4	2	2
5	5	5	2.5	2.5
6	6	6	3	3
7	7	7	3.5	3.5
8	8	8	4	4
9	9	12	4.5	6
10	10	16	5	8
11	11	20	5.5	10
12	12	24	6	12
13	13	28	6.5	14
14	$\geq$14	$\geq$32	$\geq$7	$\geq$16

Table 5: True ToT value in bunch crossing (BX = 25 ns units) for each output ToT 4-bit code, depending on speed (40 or 80 MHz) and compression (4 bit or 6-to-4 bit) settings. Always the low edge of the true ToT bin is shown. For example code 3 having a true ToT low edge of 3 means the true ToT was at least 3 bunch crossings and at most x, where x is the true ToT low edge of the next code (4 in this case). The last bin (code 14) has no high edge and includes all overflows. Code 15 means "no hit" and should never be seen because unhit pixels are internally suppressed.

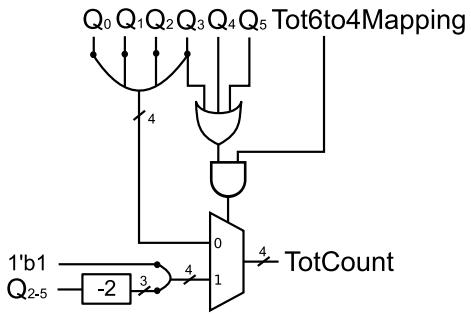


Figure 45: ToT 6-to-4 bit mapping schematic.

once the conversion is finished, indicated by a trailing edge pulse or by the counter reaching max count. Which of the memories the ToT is stored in is fixed at the start of the ToT conversion by a write address from the LTR, shown as wr. addr. in Fig. 40. This is common to the 4 pixels in the pixel region.

The pixel ToT memory bank has a 4-bit output port. Which ToT memory is presented on this port is given by a select address from the LTR block (rd. addr. in Fig. 40).

7.4 Latency, Trigger and Readout (LTR) block

If any pixel in the pixel region is hit, the timestamp is stored in a memory common to the whole

pixel region. If multiple pixels fire in the same region in the same crossing, still only one memory is written. The timestamp is necessary to determine when the trigger latency for a certain event expires and the storage elements are therefore often referred to as latency memories.

Each 4-pixel region has 8 latency memories just as each pixel has 8 local ToT storage registers. Each of the 8 ToT registers of a given pixel is associated with one region latency memory (hard-wired). This way, when one pixel is counting ToT it does not prevent the other pixels in the region from being hit, i.e. there is no region dead time, only single pixels have dead time while the comparator is high.

Each memory consists of a 9-bit buffer and comparator, as shown in Fig. 46, which is a low power solution compared to local latency counters. The 9-bit memory stores the value of a global bunch crossing counter (BCID) distributed from the chip bottom (9 bits translates to a maximum trigger latency of 2^9-1=511 bunch crossings or 12.8 μs). Every subsequent bunch crossing, this value is compared to a delayed bunch crossing counter (BCID request), also global, that is delayed by the trigger latency relative to the BCID. Both are Gray counters so that only one of the 9 bits changes every bunch crossing. If the event is triggered, the same memories are recycled to store the trigger ID value.

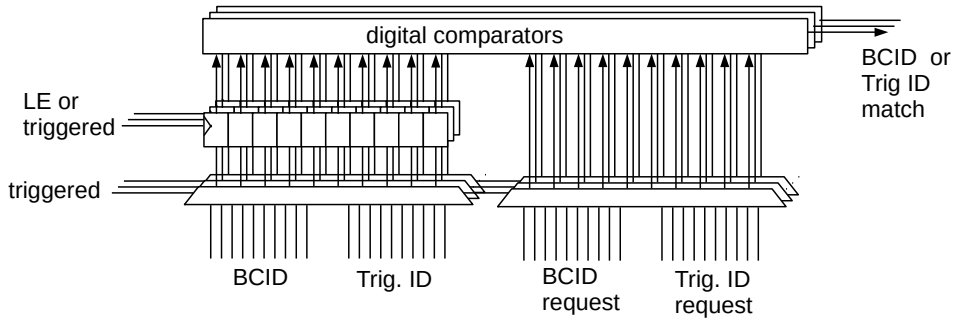


Figure 46: Latency memory block diagram. The same memory and comparators are used to store BCID values or trigger ID values, as there is never a case when both need to be stored at the same time. Only 3 instances of each circuit are shown in the figure, but in reality there are 8 9-bit memories and comparators in the region.

Each memory cell also contains a 2-bit state register to identify whether the memory is idle, triggered, to be read out or to be cleared. The latter is only used in two-level trigger mode or for event truncation. Two-level trigger mode is a prototype functionality that was not selected for use in the final ATLAS or CMS detectors. Therefore, while partly present in RD53C, it has not been fully supported. The memory is in idle state until a leading edge (LE) signal arrives. With each new LE pulse, a new memory location is written. When a location reaches the programmed latency (the BCID and BCID request values match) the presence of a trigger is checked. If no trigger is present, the latency buffer and associated ToT memory are released (marked available). If a trigger is present, then the ToT memories are marked triggered (not available) and a trigger ID is stored to label the hits for later readout.

When a region hit is selected by a trigger, the latency buffer mechanism is reused for queuing

the hit for readout. The stored BCID value is overwritten with a trigger ID value, which is compared to a trigger ID request value. When the trigger ID request matches the stored value, the LTR will hold the read token and select ToT data to be placed on the output. The read token travels from one region to the next to scan a core column for data. The generation of the trigger ID is explained in Sec. 9.3.

The Precision ToT modules in the chip bottom reuse the region latency memory mechanism and readout logic. Sec. 13.7 may give further insight about the region operation.

7.5 Pixel Addressing

The pixel addressing is hierarchical, first in cores (like postal codes) and then regions within a core (like the street address). This structure is shown graphically in Fig. 47.

For writing pixel configuration values, the basic unit which can be addressed is a pixel pair (see Fig. 47). This is achieved by writing to two registers of the chip global configuration. The Core_Col and Core_Row values are preserved in the global configuration registers 1 and 2, respectively, but the four Region_in_Core are divided between the two registers as follows. Additionally, there is a Pair_in_Region bit need (17 total bits instead of 16), because configuration is written in pixel pairs rather than pixel regions or individual pixels.

```
Register 1= [7:2]=Core_Col, [1]=Region_in_Core[0], [0]=Pair_in_Region
Register 2= [8:3]=Core_Row, [2:0]=Region_in_Core[3:1]
```

In this way, Register 1 identifies the pixel pair column address, while Register 2 contains the pixel row address.

In the RD53C data output (Sec. 10.4) the basic unit is instead a quarter core, which contains two rows of 8 pixels. Thus, a core from Fig. 47 is vertically divided into four quarter cores, with the numbering of the quarter cores as shown in Fig. 48. A binary tree compression scheme is used to encode the address of each pixel in a quarter core, as described in details in Sec. 10.4. Fig. 48 also shows an example of the compressed bit codes for all 16 cases of a single hit pixel in the quarter core.

7.5.1 25 μm $\times$ 100 μm pixels

While for 50 μm $\times$ 50 μm sensors the chip pixel numbering will carry over unchanged to the sensor, for 25 μm $\times$ 100 μm sensors a mapping is needed to know which sensor pixel is connected to which chip channel. This mapping is determined by the sensor metalization and there are two possible mappings as shown in Fig. 49.

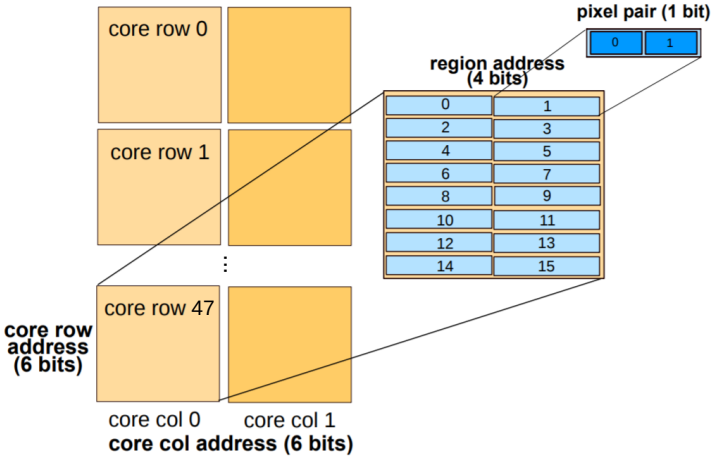


Figure 47: Pixel core addressing scheme, down to pixel pairs for pixel register configuration.

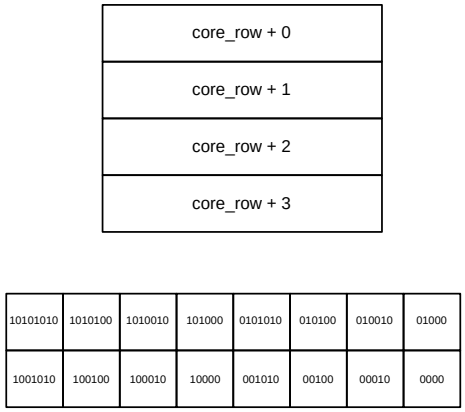


Figure 48: Numbering of quarter-cores for readout (top) and compressed binary code for each single hit pixel (bottom).

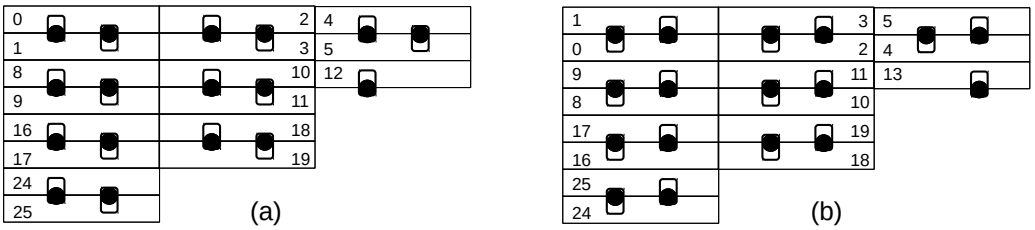


Figure 49: Two options (a and b) for mapping of 25 μm $\times$ 100 μm pixel sensors to the core pixel address. Which option is correct is determined by the sensor metalization. The filled circles represent the bump bond locations while the open rounded rectangles extend from each bump to the center line of the sensor pixel served. The top left corner of an 8 by 8 pixel core is shown.

8. Commands and Configuration

RD53C is fully controlled with a 160 Mbps differential serial input stream with a custom, DC-balanced encoding described in Sec. 8.2. The differential receiver circuit is described in Sec. 8.1. The received signal, without any processing, can be optionally repeated on the general purpose differential outputs (Sec. 13.1). The command input also contains an activity detector that will cause a reset when the rate of transitions falls to very low value (Sec. 3 and 14). A Clock and Data Recovery circuit (CDR) recovers the input bitstream and also produces the internal clocks for the chip, based on the transitions on input stream, as described in Sec. 14. A dedicated command (PLL_LOCK) equivalent to a clock pattern is provided to ease locking the internal phase locked loop (Sec. 14).

8.1 Receiver Circuit

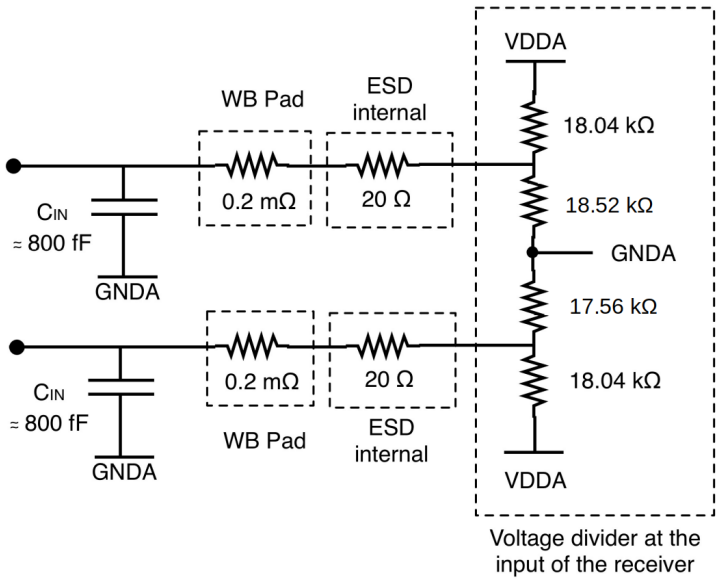


Figure 50: Equivalent circuit for differential receiver input.

The CMD receiver is implemented as a differential amplifier with a rail-to-rail input stage. The inputs are connected to an on-chip resistor bias network in the kΩ range, which allows the receiver to be ac-coupled to the serial input stream. The resulting input common-mode voltage is VDD_PLL/2 (600 mV for nominal 1.2 V supply). The bias network also adds a small offset voltage to the differential input signal to keep the receiver in a static input state in case of a broken signal connection. A termination resistance is not implemented in the CMD receiver to allow multiple RD53C chips to be connected to the same CMD line (a multi-drop configuration with one termination at the end). The differential receivers for the data aggregation inputs (Sec. 11) use the same rail-to-rail input stage . The input impedance of the receiver, together with the ESD protection and wire bond pads, have been simulated with extracted parasitics. The capacitive and resistive contributions are shown Fig. 50.

8.2 Command Protocol

The input stream is a continuous sequence of commands. All commands are built in 16-bit frames made out of two 8-bit symbols. As the bitrate is 160 Mbps each frame spans four periods of the 40 MHz bunch crossing clock. Commands that are one frame long (four BX clocks) and are called short commands (Sec. refsec:short-commands) of which Triggers are an example. Frames are interpreted one at a time and short commands are executed immediately, while long (multi-frame) commands (Sec. 8.2.2) are executed after their last frame is received. Long commands have the property that they can be interrupted by short commands without the need of restarting the interrupted command. This gives the ability to send Trigger commands (which are short) whenever needed, and to send long commands during data taking without worrying if trigger might be coming.

The command input is intended to be shared by multiple chips (multi-drop). Commands can be *broadcast*, in which case all chips sharing the command input will execute them, or *addressed*, in which case only the chip with the selected address will execute it and all other chips receiving it will ignore it. A chip can have one of 16 possible chip ID values (set by 4 wire bonds to ground overriding internal pull-up resistors). The first frame of addressed commands consists of an 8-bit symbol identifying which of the 7 commands it is, and a data 8-bit symbol specifying a chip ID. Addressed commands can also be sent in broadcast mode by specifying a chip ID value greater than 15. A chip that receives a command not broadcast or addressed to it will still process it (so as not to produce "unexpected data frame" errors), but will not execute it. The PLL_LOCK, Sync, and trigger command are always broadcast, while all others are addressed.

Each 16-bit frame is exactly DC balanced. DC balance is needed for A/C coupling, reliable transmission, and clock recovery. The symbols used also provide error detection [3]. There is a unique sync frame (used to perform frame alignment as explained in Sec. 8.3), plus 3 kinds of TTC (Trigger, Timing and Control) frames: trigger, command, or data. TTC frames contain two 8-bit symbols which are themselves DC-balanced. Furthermore, symbols that begin or end with three or more 1's or 0's are not used, resulting in a maximum run length of 4, except for the sync frame which has a run length of 6. The valid symbols and commands are given in Tables 6, 7, and 32. There is one sync frame, 7 non-trigger commands, 15 trigger symbols allowing the encoding of 15 trigger patterns (Tables 6, 7), and 32 data symbols allowing the encoding of 10 bits of content per data frame or 5 bits of chip ID per command frame (Table 32). All valid symbols are allowed to be used as trigger tags in the trigger frame; thus there are 54 possible tags (see Sec. 9). A single bit flip always results in an invalid symbol (formally, all symbols are separated by a Hamming distance of 2).

RD53C interprets the protocol in three phases (which will be transparent to the user): Initialization 8.3, Data Transmission 8.4 and Decoding 8.5. The decoding timing and exception handling are covered in Sec. 8.6

[3] All these properties could have been obtained with 8b/10b encoding, but the 10-bit frame length of 8b/10b would have required 200 Mbps link speed in order to maintain an integer number of bunch crossings per frame, as needed for synchronous triggering. The 160 Mbps bitrate of the RD53C custom protocol makes for better transmission on low mass cables and can be directly driven from GBT e-links.

Command	Encoding		(T)ag, (A)ddress or (D)ata 5-bit content					
Sync	1000_0001	0111_1110						
PLLlock	1010_1010	1010_1010						
Trigger	tttt_tttt	Tag[0..53]						
Clear	0101_1010	ID<4:0>						
Global Pulse	0101_1100	ID<4:0>						
Cal	0110_0011	ID<4:0>	D<19:15>	D<14:10>	D<9:5>	D<4:0>		
WrReg(0)	0110_0110	ID<4:0>	0,A<8:5>	A<4:0>	D<15:11>	D<10:6>	D<5:1>	D<0>,0000
WrReg(1)	0110_0110	ID<4:0>	1,xxxx	xxxxx	N×(D<9:5>	D<4:0>)		
RdReg	0110_0101	ID<4:0>	0,A<8:5>	A<4:0>				
Read_trigger(*)	0110_1001	ID<4:0>	00,T<7:5>	T<4:0>				

Table 6: List of protocol commands/frames and address or data fields associated with each. Unused padding bits are indicated by "0". Double vertical lines denote frame boundaries. tttt_tttt is one of 15 trigger commands (Table 7). The before-encoded bit content of chip ID, Address or Data is shown. These are all encoded as 8-bit data symbols (Table 32). (*) Read_trigger is a legacy command and should not be used in RD53C, as the trigger mode requiring it has been deprecated.

8.2.1 Short Commands

PLL_LOCK (broadcast only):

This command allows a clock pattern to be sent to the chip without any action being executed by the command decoder. The clock pattern is needed to efficiently lock the Phase Locked Loop (PLL) to the correct frequency at start of operation (Sec. 14). Once locked, the PLL no longer needs a perfect clock pattern and regular commands and sync frames can be sent. This command can also be used as an idle when there is nothing to be sent during normal operation. This is equivalent to a No Operation (NOOP) command in many processors, but we do not use that terminology here. It repeats the same 8-bit symbol twice to produce a clock pattern (Table 6).

Sync (broadcast only):

The Sync is the only command where the two 8-bit symbols used are not themselves DC balanced (both together the 16 bits are DC balanced). This is what makes it unique and allows it to be recognized for frame alignment.

Clear:

Clears the entire data path. All pending triggers and stored hits will be erased. This command can be used when every chip receiving it has dedicated readout link(s). However, when using data merging to read out multiple chips on a single link, the command should not be used and should instead be replaced by a Global Pulse command. See Sec. 15.

Global Pulse:

The global pulse command sends a single pulse with a duration of N bunch crossings, where N is the value of the 9-bit register GlobalPulseWidth (Table 22). The value $N = 0$ is treated like $N = 1$. The global pulse can be routed to different places of the chip and has many

uses. It can provide reset signals, control the ring oscillators, the ADC, etc. The global pulse routing table is 24.

Trigger (broadcast only):
Because one 16-bit frame spans 4 LHC bunch crossings, the trigger command must specify a 4-bit map indicating which of the 4 bunch crossings are actually triggered; hence 15 trigger patterns. The triggering is synchronous, and therefore trigger frames must be sent at specific times. The second symbol in a trigger frame can be any legal symbol and is interpreted as one of 54 possible 6-bit tag bases to identify the trigger(s) in later readout (see Sec. 9.2). The mapping from symbol to tag base number is given in Table 34. The trigger tag will be returned with the data corresponding to that trigger (See Sec. 10).

Symbol Name	Encoding	Trigger Pattern	Symbol Name	Encoding	Trigger Pattern
			Trigger_08	0011_1010	T000
Trigger_01	0010_1011	000T	Trigger_09	0011_1100	T00T
Trigger_02	0010_1101	00T0	Trigger_10	0100_1011	T0T0
Trigger_03	0010_1110	00TT	Trigger_11	0100_1101	T0TT
Trigger_04	0011_0011	0T00	Trigger_12	0100_1110	TT00
Trigger_05	0011_0101	0T0T	Trigger_13	0101_0011	TT0T
Trigger_06	0011_0110	0TT0	Trigger_14	0101_0101	TTT0
Trigger_07	0011_1001	0TTT	Trigger_15	0101_0110	TTTT

Table 7: List of trigger symbols used to encode the 15 possible trigger patterns spanning four bunch crossings. Note there is no 0000 pattern as that is the absence of an trigger. The Trigger_01 (000T) means that the first bunch crossing of the trigger window is meant to be readout, and the extended tag returned will have 00 following the supplied tag base.

8.2.2 Long Commands

Cal (Calibration Injection):
The same command is used for both analog and digital injection. Whether injection will be analog or digital is decided by global configuration register CalibrationConfig, but the Cal command produces the same output regardless. To understand the Cal command it is necessary to understand how the calibration injection circuit works. Therefore, the description of the command was given in Sec. 6.

WrReg(0) (Write Register, single):
The WrReg command has two modes: single write and multiple writes to register 0. The command frame is the same and the distinction between single and multiple is made by the first bit of the payload (0=single, 1=multiple). The WrReg(0) or single has 9 bits of Address and 16 bits of Data. Up to 512 16-bit wide registers can be addressed, but not all 512 possible register addresses are used. If an attempt is made to write to an unused address, the command will do nothing and no warning will be generated. The register memory map is given in Table 22. This command does not produce any output from the chip.

WrReg(1) (Write Register, multiple):
The WrReg command has two modes: single write and multiple writes to register 0. The command frame is the same and the distinction between single and multiple is made by the first bit of the payload (0=single, 1=multiple). The WrReg(1) or multiple must have the address value set to 0. It can only be used to initiate multiple writing to register 0. Register 0 is a virtual register called PIX_PORTAL, used to write and read pixel configuration (Sec. 8.8).

Following a WrReg(1) command, one may send data frames to the chip (as many as desired) without any preceding command. All these data frames will be written to the PIX_PORTAL (register 0). This must be done in conjunction with auto-increment (see Sec. 8.8). This permits very efficient transfer of data to the PIX_PORTAL. The write multiple mode remains in effect until a new long command is received (short commands will be executed and not end the WrReg(1) command mode). Note that the chips not addressed by the WrReg(1) command will still recognize the multiple write mode and will therefore not issue "unexpected data frame" warnings, but they will not write the data to their register 0. The placement of the 10 bits from each data frame into the 16 bits of register 0 is described in Sec. 8.8

RdReg (Read Register):
This command has 9 bits of address and no data. It initiates the readout of the addressed register. Address 0 is special: it is the the pixel register as described in Sec. 8.8. The 16-bit register value is returned in the data stream as described in Sec. 10. Not all 512 possible register addresses are used. If readback of an unused address is requested, the data value returned will be 0, the address returned will be the requested (non-existent) one, without any warning generated. The register assignment list is given in Table 22.

RdTrig (Read Trigger):
This command has an 8-bit extended tag value. In two-trigger mode, it selects a previously received tag for readout. It is not useful in single trigger mode. See Sec. 9 for details.

8.3 Command Protocol Initialization

Until the PLL is locked and produces a stable chip clock, the command decoder will be in its reset state. During this period, PLL_LOCK frames should be sent to the chip. The transitions in the string of PLL_LOCK frames will allow the clock recovery circuit to lock to the correct 160 MHz frequency. The user does not know when the PLL has locked, but simply sends PLL_LOCK frames for a long enough time that the lock cycle is surely completed (see Sec. 14). For debugging, the PLL lock condition can be observed in the recovered CMD output of the general purpose LVDS, which is a default output (see Sec. 13.1). At this point the protocol initialization begins. Before any command decoding, the input bitstream is processed by the Channel Synchronizer circuit (Fig.51), and the initialization correctly sets up this circuit.

The sync pattern (Table 6) can not be produced through any combination of TTC frames and therefore can be searched for to lock the correct frame boundaries (the search procedure is explained in the next paragraph). Sync frames must be sent at the start of operation so that the framing can be locked (this different from PLL lock!). It is mandatory to send one sync frame in every 32 frames or so in order to maintain lock or allow the command decoder to re-lock if

lock was lost. If no sync frames are received in a long time frame lock will be declared lost and the command decoder will stop interpreting commands until a new lock is acquired. Typically at the start of operation (power up) there are no commands or triggers to immediately send, and so sending a large number of sync frames to ensure initial lock is not a problem. The channel synchronizer lock is available in the chip status CMOS output (Sec. 13.1).

Using the 160 MHz recovered clock, the channel synchronizer will search for sync symbols and count each valid appearance of this pattern in 16 separate channels (one channel for each possible frame alignment). When the count for one of the channels, i, reaches a threshold N_{lock}, *sync lock* is declared as acquired, channel i is adopted as the correct channel, and the count of the remaining 15 channels is reset. The value N_{lock} has default value of 16 and can be changed in configuration register ChSyncConf (Table 22). At the start of transmission the command decoder will not interpret any commands until it has received N_{lock} Sync commands. Thus one should begin transmission by sending at least N_{lock} Sync commands. The 40 MHz bunch crossing clock is generated as the bit pattern 1100110011001100 aligned to channel i. Thus there are 4 bunch crossings with a fixed phase relationship to the sync frame, which can be labeled $BXa..BXd$. The counting of sync sequences continues in all the channels, but every new sync sequence detected on the lock channel i resets the count for all the other channels. If the count for a channel that is not the lock channel ever reaches a threshold of $N_{lock}/2$, lock is declared lost, and a new sync lock is acquired on the first channel that reaches the locking threshold N_{lock}. This allows for continuous channel monitoring and automatic sync lock as long as enough sync symbols are transmitted. Additionally, if zero sync frames are received in the lock channel within 64 frames (regardless of other channels), lock will be declared lost and no further commands will be decoded until a new lock is acquired. This value is hard-wired and cannot be changed. This is useful to prevent prolonged, random input due to an upstream exception from corrupting the chip operation, but makes it mandatory to regularly send Sync symbols.

8.4 Command Protocol Transmission

During transmission a correct sequence of commands is sent to control the chip. Trigger frames are sent at specific times, and the “space between trigger frames" is filled with commands (including the required Syncs). Long commands are decoded regardless of intervening short commands. The PLL_LOCK command can be used as an idle frame, as it has the most transitions and will therefore best maintain PLL operation. Sync commands can also be used as idles, since they must be sent periodically anyway, but they have the fewest transitions, so are not ideal for maintaining PLL lock. The best approach is therefore to always send Syncs every 32 frames and PLL_LOCK commands in between if and when there is nothing else to send.

8.5 Command Protocol Decoding

The data bits recovered from the locked channel are fed to the Command Decoder as shown in Fig. 51. In the absence of a sync lock, nothing is fed to the command decoder, so until a lock happens no commands will be interpreted. The locked condition guarantees that the bits fed to the command decoder are correctly aligned with the 40 MHz bunch crossing clock. Protocol consistency is ensured by checking that the decoded frames are valid and also that they match what is expected (analogous to checking both spelling and grammar). The 16 bits are fed to the command

decoder with a parallel bus. In case of correct detection, the indicated action is performed according to the command type and Chip ID. All symbols are always checked and decoded, even if they follow a Chip ID that does not match the wire bonded ID. However, the Command Decoder will act on the rest of the chip only if the command is a trigger, if decoded Chip ID matches the wire bonded ID, or if the decoded broadcast bit is 1 (the PLL_LOCK command is not addressed, but has no internal action- no operation). The detection of an invalid symbol is handled differently depending on the frame and expectation (current state). The handling of exceptions is shown in Table 8.

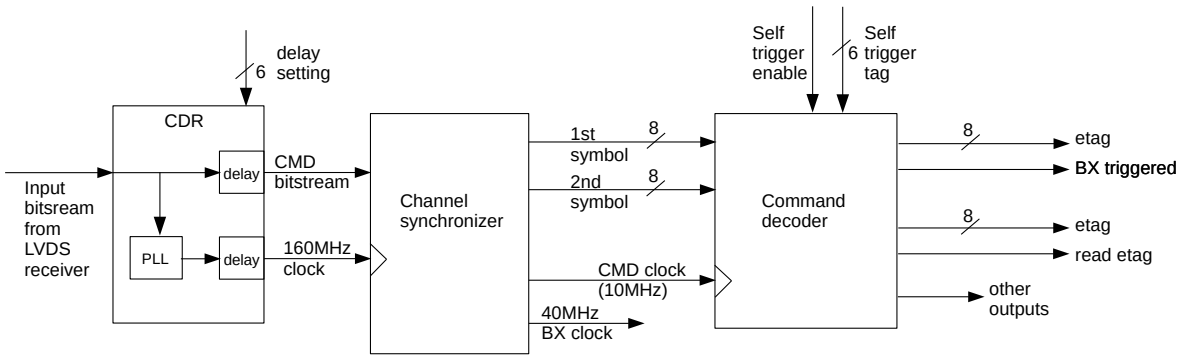


Figure 51: Clock and command recovery and decoding path from chip input to internal signals, showing trigger pluses and tags in particular. Other outputs of the command decoder, such as global register address and write signal, not shown. 16-bit Command patterns are successively loaded into the Command Decoder with the correct frame alignment as determined by the Channel Synchronizer.

8.6 Command Protocol Timing

The decoded commands are executed 25 ns after the end of the last frame of the command data. "Executed" means that the outputs of the Command Decoder block in Fig. 51 change state, which happens on a rising edge of the beam clock. In many cases the execution is instantaneous (outputs change state and that's it), but the Trigger, Cal and Global Pulse commands have a delay and duration. The trigger command sends 1 to 4 pulses in 4 consecutive beam clock cycles, and thus is completely finished before a new command can be completely received (since 1 frame is 4 beam clock cycles). The Cal and Global pulse commands can occupy their respective output lines (CAL_edge, CAL_aux, and Global_pulse) for many clock cycles. A new Cal or Global pulse command should not be sent before the prior such command is complete (up to the DAQ to ensure this), but any other command can be sent and will be executed normally.

8.7 Global Configuration

The global configuration is stored in 16-bit registers which are accessed like a RAM with the write and read register commands of Table 6. Each register has a default value that is provided as

Frame received	Frame Expected	Error/Action
invalid, data	data	Aborted command
data, invalid	data	Aborted command
invalid, invalid	data	Aborted command
invalid, data	not data	Lost trigger
invalid, invalid	not data	Corrupted frame
invalid, sync	any	Corrupted sync
sync, invalid	any	Corrupted sync
invalid, command	any	Execute with warning
command, invalid	any	Execute with warning
trigger, bit-flip (*)	any	Execute w/tag base 54
trigger, invalid (*)	any	Execute w/tag base 55
command, command	data	Ignored command

Table 8: Command Decoder response to invalid or unexpected symbols. (*) bit-flip refers to an 8-bit pattern produced from flipping a single bit in a valid symbol, while invalid references to any other invalid 8-bit pattern.

explained in Sec. 3.1.1. The main table of register names, content, and default values is given in Sec. 16.2.

8.8 Pixel Configuration

Each pixel has 8 bits of local configuration as detailed in Table. 23. From the point of view of the write and read register commands, each pixel is seen as one half of one configuration data register. All pixels are paired as shown in Sec. 7.5.

The 8 pixel bits are divided into 5 TDAC bits (threshold tuning bits) and 3 enable bits (also known as mask bits). These two types of bits can be written together or independently (always for two pixels at a time). Thus one can choose to write all 8 bits at once, only the 5 TDAC bits, or only the 3 enable bits. The single write register command (WrReg(0)) of Table 6 always writes al 8 bits of both pixels, where the 16 bit data frame is subdivided as follows:

Single Write: left-pixel(TDAC[15:11]HitBus[10]InjEn[9]Enable[8]),
right-pixel(TDAC[7:3]HitBus[2]InjEn[1]Enable[0])

The multiple write register command (WrReg(1)) instead writes the mask bits or the TDAC bits depending on the Mask or TDAC bit of global configuration register PIX_MODE. The mapping from 10-bit data frame to two pixel TDAC or mask bits is as follows:

PIX_MODE[1] = 0: unused[9:8], right-pixel-mask[7:5], unused[4:3], left-pixel-mask[2:0]
PIX_MODE[1] = 1: right-pixel-TDAC[9:5], left-pixel-TDAC[4:0]

Internally, the writing and reading of configuration values from the pixels uses an addressed bus to every 2×1 pixel pair. All reading and writing is done two pixels at a time in a given column of 4-pixel regions. (See Sec. 7.5 for address encoding). However, multiple core columns can be

written in parallel, while readback can only take place from one pixel-pair column at a time. There are thus two write modes, single pixel-pair and broadcast, while read is always single pixel-pair.

The write and read operations are controlled by three global registers, the REGION_COL, REGION_ROW, and PIX_MODE configuration registers. The pixel data is written into or retrieved from global register 0 (PIX_PORTAL) with the normal write and read register commands (see Sec. 8.2). This is a virtual register acting as a portal to whatever pixel pair is pointed to by the column and row config registers (called PIX_PORTAL). The row register has a special feature called auto increment (Auto Row), which reduces the number of commands needed to fully configure the chip. This mode is enabled by a configuration bit and increments the row register value after every write or read operation to PIX_PORTAL.

The typical pixel matrix configuration write sequence, using the write single register command, is given in Table 9. Note that this takes 77200 (73008) commands for ATLAS (CMS) chips to accomplish. These numbers should be multiplied times 4 to obtain number of frames, and each frame takes 100 ns to transmit. If one is only configuring a chip, it will therefore take about 30 ms. For the case of configuring during data taking (called trickle configuration), much of the command bandwidth will be taken up by trigger commands, and configuration will therefore take longer. The worst case is two-level trigger operation with 4 MHz L0 + 1 MHz L1 trigger operation. This will use up 60% of the command bandwidth. We should also remember that 6% of the command bandwidth must be used to send periodic Sync commands. with only 34% of the command bandwidth available, 77200 Write Register commands will take 88 ms instead of 30 ms. With some DAQ overheads we assume 100 ms. So for a 4-chip module trickle configuration in the worst case will take 400 ms. (If more chips share the same command line it will take proportionally longer). Writing a uniform (all pixels the same) configuration is 50 (54) times faster for ATLAS (CMS), because each Write Pixel command can write to all core columns Table 10. Alternatively, using the multiple instead of single Write Register command means one frame instead of 40 frames per write, which will reduce the above 88 ms to 22 ms (100 ms for a 4-chip module in worst case of trickle configuration). The readback of the pixel configuration for the whole matrix can proceed exactly as shown in Table 9, substituting the Read Register command instead of Write Register. This can be carried out in broadcast mode to any number of chips in parallel, so will always take 50 ms (half as much as writing a single chip because the read register command is two frames instead of four).

Writing or reading an individual, arbitrary pixel pair follows steps 1-3 of Table 9. For calibration operations it is often required to write only the mask bits many times to shift a pattern through the matrix, leaving the TDAC bits alone. This can be done with broadcasted commands (same mask for all chips even though the TDACs are different), and it must use the write multiple command as the write single command always writes all the 8 configuration bits per pixel. Writing masks to a single pixel at-a-time will take 77200 write operations as in Table 9, but one frame per pixel write instead of four (still four frames per write for steps 1 and 2), resulting in 78400 frames, which takes 8.3ms if all the command bandwidth (minus 6% for Syncs) is used, or 23 ms in the worst case of trickle calibration. Writing one row at a time will take 1/50 of this per mask, regardless of the number of chips, as it is done in column broadcast mode using broadcasted commands. It can even be faster if not all rows need a new mask each time.

When used in a radiation environment it is possible to write to an non-existing pixel address at the end of a configuration operation. This will make pixel configuration less sensitive to accidental

Step	Command	Address	Explanation
1	Write_Register	column and mode config	set columns 0-1 and auto row mode
2	Write_Register	row config	set row 0
3	Write_Register	0	config first 2 pixels
4	Write_Register	0	config for next row 2 pixels
386	Write_Register	0	config for last row 2 pixels in cols 0-1
387	Write_Register	column and mode config	set columns 2-3 and auto row mode
388	Write_Register	row config	set row 0
389	Write_Register	0	config for next row 2 pixels
77200	Write_Register	0	config last 2 pixels in chip

Table 9: Sequence to write an arbitrary pixel configuration to ATLAS size chip using write register single commands. Each column pair takes 386 commands, times 200 column pairs leads to 77200 commands. For readback replace Write_Register 0 with Read_Register 0 commands.

Step	Command	Address	Explanation
1	Write_Register	column and mode config	set broadcast, cols. 0-1, and auto row mode
2	Write_Register	row config	set row 0
3	Write_Register	0	config all pixels, first row in cols 0-1
4	Write_Register	0	config all pixels, second row
386	Write_Register	0	config all pixels, last row
387	Write_Register	column and mode config	set broadcast, cols. 0-1, and auto row mode
388	Write_Register	row config	set row 0
389	Write_Register	0	config all pixels, first row in cols 2-3
1544	Write_Register	0	config last 2 pixels in cols 6-7

Table 10: Sequence to write a default (all pixels the same) configuration for ATLAS size chip. Only the first core column (columns 0-7) are written because all core columns will be "CC-ed" in parallel.

SEU/SET caused overwriting a pixel register.

9. Trigger Processing, Tags, and Data Flow

While the RD53C design contains two trigger modes for historical reasons, known as single level and two-level, the two-level trigger function has been deprecated, has not been fully verified, and should not be used. Only the single level trigger, which is the default and is fully verified, is described here. For a description of two-level trigger refer to the RD53B manual.

A simplified description of the chip triggered readout is as follows.

1. A trigger command is received, including an identifier called tag,
2. The tag (see Sec. 9.2) is stored in a trigger table (Sec. 9.3). and the hits from the appropriate bunch crossing are associated with this tag (but left in the pixel matrix),
3. The hits are read from all core columns in parallel, and then assembled into whole events at the chip bottom, along with the tag. Whole events (tagged) are placed in streams and sent to the Aurora encoder for output.
4. The processed tag is erased from storage and is now available to be used again for another trigger.

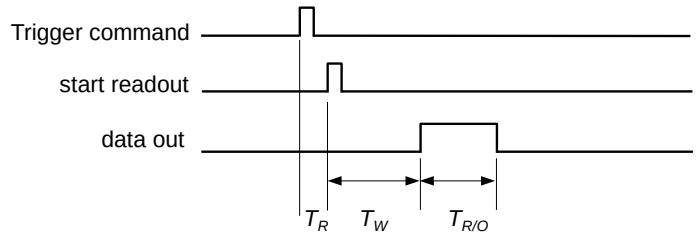


Figure 52: Timing of trigger to data readout.

Fig. 52 shows the timing from a trigger to the completion of data readout. The time to start of read (T_R) is a fixed delay from processing the trigger command. This is followed by a variable delay before data for that trigger come out of the chip, T_W, because the trigger must wait its turn behind prior triggers. This is a simple queuing wait time. The event readout time $T_{R/O}$, scales with the hit occupancy (see Sec. 9.1).

The hit data flow can be understood as a three stage process. The core column readout, the aggregation of data for each event from multiple core columns, and the Aurora encoding, serialization and output. Each stage pulls data from the previous stage when it needs it. Thus buffers are generally not empty. When buffers are empty idles are inserted into the output stream.

Each core column works as a self-contained unit with a small amount of storage at the bottom of the column. Its job is to pull data from the pixels to try to fill its bottom of column buffer (it pauses when buffer is full), regardless of whatever else is going on in the chip. All core columns do this in parallel and independently. Each core column has its own ordered list of triggered BCIDs to read out. Even if at T=0 all these column lists are identical, that will not be the case for long, as each column will work through its list at its own pace given by what hits it contains. The encoding

of hit maps and ToT is done in the column readout. Thus, the bottom of column buffers contain encoded bits, not individual hits.

The event building proceeds one triggered and selected-for-readout BCID at a time. It pulls the data for that BCID from the column buffers that contain any. Many columns may not contain any hits for that BCID and are skipped. The event building BCID will always be the first one present for columns that have hits, because columns and event building process triggers in the same order. The column numbers for a given event do not have to come out in numerical order- it depends on which column is ready first. A given column address could also appear more than once if it contains many hits. The event tag, and stream markers are added by the event building stage. The built event data are placed in a buffer to make them available to the Aurora encoder. More details are given in Sec. 9.5.

The Aurora encoder pulls data from the event buffer, performs the Aurora formatting, idle insertion, etc. The contents are serialized onto the output lanes. Data merging modifies how this stage works. The following subsections give a more detailed description of the trigger processing and book-keeping. Technical details of the bottom of chip data flow are given in Sec. 9.5.

9.1 Pixel Matrix Processing and Wait Time

The pixel matrix operates in steps of the 40 MHz beam crossing clock (BX). Within the matrix each BX is triggered or not triggered based on the state of a trigger signal (high is triggered, low is not). Thus a trigger is a one BX long pulse on this trigger signal. Trigger pulses are normally issued by the command decoder in response to commands received. The trigger command path, from chip input to internal trigger pulses, is shown Fig. 51. Note that the internal BX clock is generated by the channel synchronizer based on the frame alignment of the input control stream (see Sec. 8.3). An individual chip phase adjustment, in 1/(1.28 MHz) steps, is introduced by the clock and data recovery circuit (CDR). Thus, each chip can be individually "timed in" to the bunch crossings.

Each of the 15 trigger commands of Table 7 generates a different pattern of pulses spanning four BX's. Trigger pulses can also be generated by the internal self trigger source (Sec. 9.4). The command decoder arbitrates the trigger sources, with trigger commands always having priority.

In the matrix, each trigger pulse marks data as triggered and associates it with a trigger identifier (ID), but *does not* initiate readout. The readout of data marked by a trigger ID is initiated later, after most of T_W in Fig. 52. In addition to queuing wait time, T_W Contains fixed delays, including the 3 BX token transit to retrieve data from the pixel matrix, another 3 BX for column hit data encoding, the Aurora encoding, etc. The sum of all these fixed delays defines the minimum possible T_W and is 31 BX. The readout time ($T_{R/O}$) is given by the number of bits being sent out times the output multi-lane bit rate, which can be up to 5.12 Gbps (4 lanes at 1.28 Gbps each). At a given trigger rate, the average $T_{R/O}$ must be less than the mean trigger period (λ), and significantly less to avoid long queuing wait time. The whole chip can be analyzed as a single server queue, which means that the wait time plus readout time $T_W + T_{R/O}$ will have a distribution like Eq. 9.1,

$$P(W > t) = \frac{T_{R/O}}{\lambda} e^{-(\lambda - T_{R/O})t} \tag{9.1}$$

It is clear from Eq. 9.1 that as $T_{R/O}$ approaches λ the total wait time diverges. The condition $T_{R/O} = \lambda$ roughly corresponds to 100% data link occupancy.

9.2 Tags

The term tag is overloaded with two meanings. In the command protocol *tag* refers to the 6-bit code received with each trigger command (more correctly called tag base). The tag base can only take on 54 values, which is the total number of DC-balanced symbols. Inside the chip and in the output data *tag* refers to an 8-bit extended tag. The two additional bits indicate which of the four BX's spanned by a trigger command the data correspond to. For example, command Trigger_04 in Table 7 with tag base value abcdef will result in one extended tag: abcdef01, while Trigger_05 will result in two extended tags: abcdef01 and abcdef11 (along with two trigger pulses).

While there are only 54 tag bases, which can lead to at most $4\times54 = 216$ extended tags, the extended tag space in the chip spans all 256 8-bit codes. The extra codes that cannot be generated from one of the 54 tag bases are used to label self-triggered events or to signal detected error conditions, as indicated in Table 11. In RD53C the main use of special tags is to label self-triggered events. Since self-triggers are generated internally in the chip, they are not constrained to the 54 tag bases, and so they are labeled with extended tags that could never result from a trigger command.

Tag values (decimal)	**Meaning**
0-215	extended tags from trigger command
216-219	Single bit-flip detected in tag symbol of a trig. command
220-223	Unrecognized tag symbol
224-255	Self-trigger tag values

Table 11: Possible extended tag values and their meaning.

The number of tag bases available is large compared to the number of triggers expected to be pending at any given time, giving the DAQ flexibility for selecting and managing the tag base value sent with each trigger to make sure an extended tag value that is already in use is never requested. Requesting an extended tag value already in use will result in the chip skipping the trigger (and incrementing the skipped trigger error counter). The simplest approach the DAQ can take is to assign a new tag base value (eg. by incrementing a counter) to each new trigger command, regardless of the command. This is "wasteful" in the sense that two different single trigger commands could share the same tag base without resulting in duplicate extended tags, and statistically most trigger commands will be single trigger commands at 1 MHz trigger rate. However, since the number of available tag bases is large, the DAQ can afford this luxury. This is because the worst case wait time for a trigger to be fully read out is simulated to be about 25 μs [5] and the Poisson probability of a random process with mean 25 (number of triggers in 25 μs at 1 MHz trigger rate) to fluctuate up to 54 or more is negligible. Therefore, the DAQ tag base counter will never come around to the same value while a given trigger is still waiting to be read out. If higher trigger rate is desired for some applications, more complex tag base selection schemes can be used.

9.3 Trigger Table

RD53C keeps track of pending and in progress triggers using a main table holding all triggers plus one dedicated table in each core column listing those triggers staged for readout, as shown in Fig. 53. The main table has 256 rows and so can hold every possible extended tag at once. The row

number is used as a trigger ID to label triggered hits in the pixel matrix. At any give time, there is a 1-to-1 correspondence between a trigger ID value and a trigger tag stored in the main table, but tags are arbitrary and user selected while trigger IDs are sequential and assigned by the chip. The trigger ID is internally Gray coded so that only one bit ever changes from one ID to the next in the pixel matrix bus distribution bus. Each trigger table row contains (1) the extended tag received with the trigger command, which is not used in the matrix, but simply stored in the main table so that it can be returned with the event data, and (2) a state for this row.

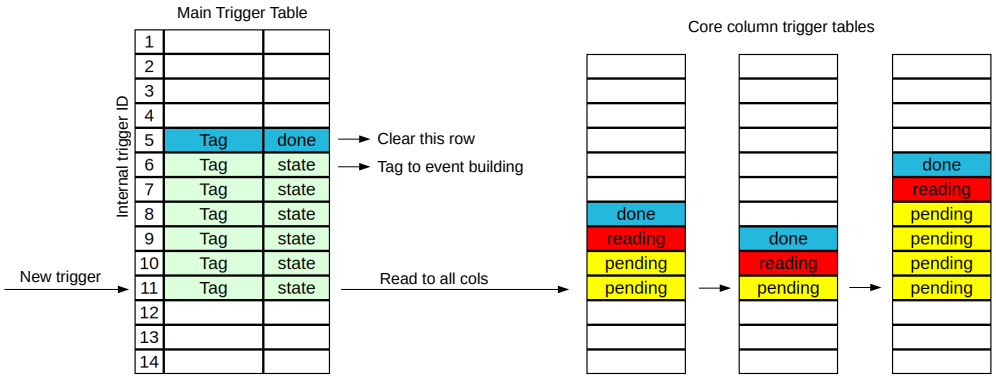


Figure 53: Conceptual diagram of trigger tables in RD53C. The row number is used as the trigger ID in the pixel matrix. The tables are circular buffers filled and emptied as explained in the text.

The column tables also have 256 rows and can be regarded as extra columns added to the main table, but their processing advances asynchronously. When a trigger arrives, it is assigned to the next Empty row in the main table and the trigger ID (row number) is sent to the pixel matrix to label the corresponding hits (see Sec. 7.4 for how the hits are time-stamped and selected in the pixel matrix). The tag is saved and row state is changed to Triggered/Read in the main table. As more triggers arrive others rows will change state to Triggered, but they will not affect this row. The new row state is now propagated to all the column tables and changes from Empty to Read in all core column tables at the same time. While all the column tables are filled in parallel this way, they are emptied independently. Each core column has it's own state machine and processing, whose job is to change rows back to Empty as fast as possible, regardless of what other columns are doing. Starting from the top down, the first non Empty row will be processed, by transitioning first from To-Read to To-Clear and then from To-Clear to Empty. The state is changed as soon as the readout or clear operations are performed on the all the hits labeled with the corresponding trigger ID (row number). Because they dispatch their non-Empty states independently from one another, each column can have a different number of rows in a given state, as illustrated in Fig. 53. Independent processing of all columns in parallel is necessary, because each core column takes at least 3 bunch crossing clock cycles to read between 1 and 8 hits (depending on hit pattern), whereas the full chip output bandwidth can only be as high as 15 hits per single clock cycle (5.12 Gbps and 10 bits per hit case).

In the main table, the state of each row with a to-Read state will change back to Empty when the event is fed to Aurora encoder. The stored tag value will be retrieved and returned in hit data stream (Sec. 10.3). Only after this point can the same tag be reused by the DAQ. Exceptions can happen, for example if a trigger arrives with a tag value that is already stored in the main table. Special tag values are reserved to mark such exceptions as covered in Sec. 9.2. It could also happen that row number is not arriving at the Aurora encoder for a very long time, preventing the row state from returning to Empty and also preventing processing of subsequent rows. To protect against this, a time-out has been implemented that keeps track of how long a row has been in the To-Read state (this uses a BCID register and comparator for every row, not shown in the figure). The timeout value is programmable (register TruncationTimeoutConf in Table 22). When the timeout is reached before the row state becomes Empty, the state is changed to To-Clear in all columns that have not yet processed the event and the state in the main table is changed to Empty.

9.4 Self Trigger Source

The self trigger functionality is a stand alone block that, if enabled, can store triggers to be processed in the trigger table. The self trigger can operate in parallel to the normal single level trigger operation from the Command Decoder, but command decoder always has priority over the self trigger. The self trigger can not operate in two level trigger mode.

The block diagram of the self trigger processing pipeline is shown in Fig. 54. The mapping of the registers found in the drawing to global configuration can be found in Table 12. In a core column there are 4 HitOr lanes to which ORs the discriminator output of the pixels (Sec. 13.4). There is a configuration bit (in the pixel register) for each pixel to activate the pixel for the HitOr. At the end of the core column each HitOr lane can be enabled or disabled via the global register HITOR_MASK_1/2/3/4.

There is one digital threshold block for each lane for each core column. The digital threshold can be disabled via HitOrDigThrEn which will also bypass any synchronization of the HitOr signal to the 40 MHz clock. Enabling the digital threshold will synchronize the HitOr signal to the 40MHz bunch crossing clock and the required threshold length can be set from 1 clock cycle up to 14 clock cycles. If a HitOr signal passes the threshold is will produce a single clock cycle pulse.

After each lane of the all core columns are ORed and fed into a large lookup table. Each entry in the table describes a unique state of all possible HitOr lane combinations. As the HitOrs are laid out in such a way that a coincidence on specific lanes corresponds to a multi pixel cluster hit in a specific direction (depends on sensor geometry).

The Pattern LUT generates a single pulse, which can be delayed up to 511 clock cycles to match it with the L0 latency. Multiple trigger pulses can enter this delay shift register. The single pulse from the pattern LUT can be elongated (multiplied) to up to 31 bunch crossings The self trigger tags are full 8-bit tags picked sequentially in the range given in Table 11, not constructed from a tag base plus two bits. The self trigger tag counter is reset by the DataPathReset signal of the Global Pulse.

9.5 Data Flow

The starting point of the data flow is hit data stored in pixel regions waiting to be read out. Such

Register Name	Bits	Field Name	Description
SelfTriggerConfig_1	[3 : 0]	HitOrDigThr	If digital threshold enabled this is the length in clock cycles the HitOr has to be active. Values 0 and 15 are allowed, but will render the self trigger unusable.
	[4]	HitOrDigThrEn	Enables digital threshold, if disabled analog (not synchronized) signal
	[5]	SelfTriggerEn	Enables (gates) output of self trigger block
SelfTriggerConfig_0	[4 : 0]	SelfTriggerMultiplier	A single trigger pulse can be elongated to cover up to 31 bunch crossings. Value 0 is valid but will render the self0trigger unusable.
	[14 : 5]	SelfTriggerDelay	Delay applied to the HitOr pulse, has to match the configured Latency such that the resulting pulse triggers the right bunch crossing. The Self trigger pipeline has in internal delay of around 12BC (depends on digital threshold).
HitOrPatternLUT	[15 : 0]		Each bit represents a unique combination of the four HitOr, this enables to only trigger of coincidence of multiple (specific) HitOrs. Note that the LSB should never be high, as it represents all HitOrs being low. 0xFFFE represents an Or of all possible combination.

Table 12: Selection of inputs to global OR operation feeding the self trigger generation.

stored hit data are labeled with a trigger ID. That labeling was carried out when the trigger arrived and all regions with hits from the BCID corresponding to that trigger were flagged (see Sec. 7).

Fig. 55 shows how data flows out of the regions and though the chip. Each core column has its own list of pending triggers (as was shown in Fig. 53) and processes that list as fast as possible, independent of all other columns. The processing pauses whenever the pending list is empty or the End Of Column buffer (EOC) is full. The hit data are retrieved one 4-pixel region at-a-time using a token that finds those region buffers matching the requested trigger. By default it takes 3 BX clocks to retrieve and encode the data from one region, but this can be increased by configuration as may be needed after logic slows down due to radiation damage. The data from 4 regions (16 pixels total) are accumulated and passed through a pipelined encoder that generates (and optionally binary tree encodes) the 16 bit hit map and discards empty ToT values. It also adds the row address, neighbor, and last hit flags. These data are placed in one row of an input EOC buffer that is wide enough to accept the maximum possible number of bits from one encoded 16-pixel region. Each row of this buffer will necessarily contain many empty bits (the max number of bits occurs very rarely). These data are then barrel shifted and packed to remove the empty bits into the output EOC buffer. The aggregation of data from different core columns can now begin.

Every 8 core columns are combined into one Data Concentrator (DC). There are 7 Data Concentrators, but the last one contains fewer than 8 core columns because the number of core columns is not a multiple of 8. Each DC runs in parallel, independently of the others. It aggregates data as fast as possible, pausing when the input buffers are empty or the output buffers are full. It processes one trigger at a time. As in the EOC, there are input and output FIFOs with a barrel shifter in between, in order to combine data from multiple columns without empty space. Data for a given

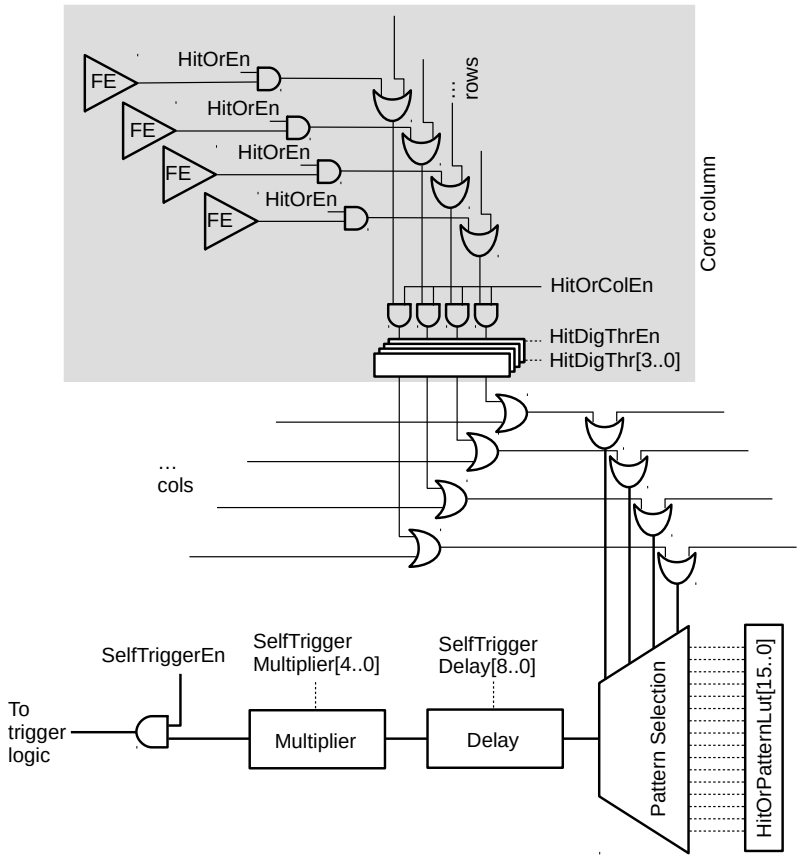


Figure 54: Block diagram of the self trigger pipeline from pixel discriminator output (HitOr) to connection to the trigger logic table.

event are pulled from one core column at a time until no more data from that event is present. Occasionally it may happen that a column has so many hits from a given event that they do not all fit into the EOC FIFOs. In this case the DC processing will visit that column more than once, and data from that EOC will appear in two different places in the same event, correctly labeled with its column number (see Sec. 10.4). Note that there is also a programmable truncation limit on the column readout (see Sec. 9.3).

In the final stage, data from the eight Data Concentrators, one at time for a given event, are pulled into the Chip Data Concentrator (CDC). This is where stream building takes place, as this is the first time that all hits from a given event are collected in the same FIFO. The tag and any other event-level information (Table 18) are added here. The total storage in the chip bottom (everything shown in Fig. 55) is 204 Kbits. As the number of bits used per hit pixel in detector readout simulations ranges between 9 and 15 (depending on occupancy and cluster distributions), that means the bottom of chip memory can hold between 14K and 20K hits. In contrast, the pixel matrix has storage for 8 hits per pixel, or 1.2M hits, so almost 100 times more (while the silicon area of the pixel matrix is only 10 times greater than the periphery).

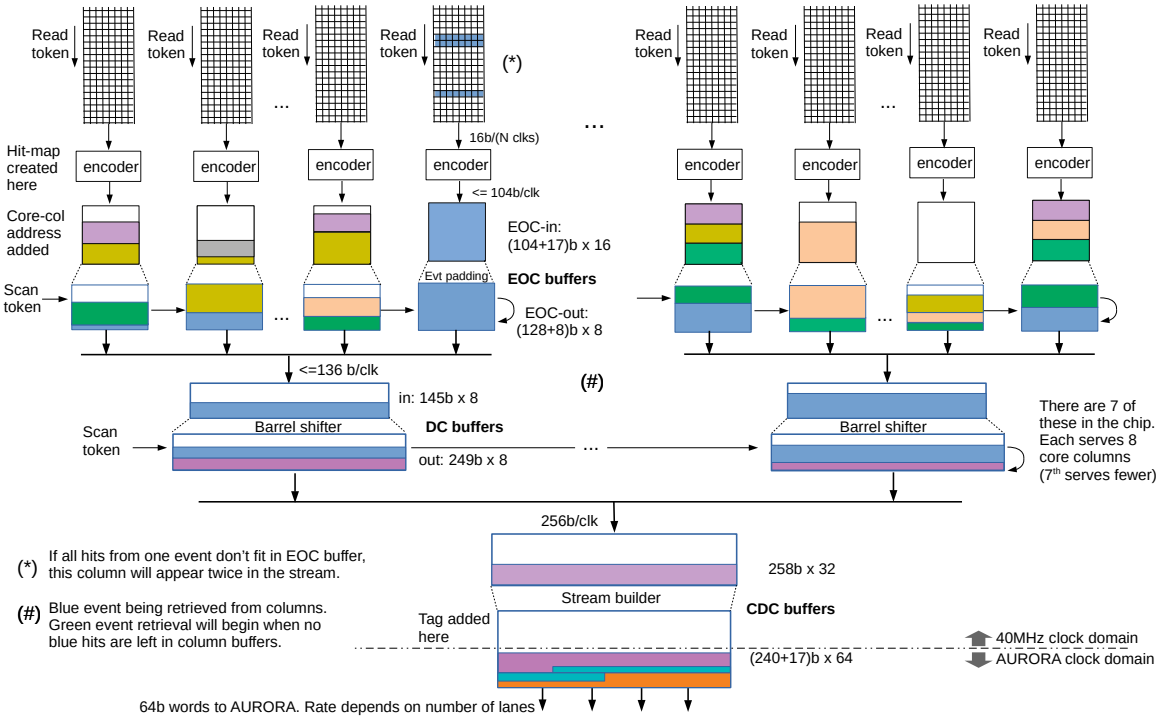


Figure 55: Schematic diagram of the data flow from the core columns to the Aurora output.

10. Data Output

The RD53C data output consists of tagged events, which enables the readout to automatically recover from transmission errors without any action from the DAQ. While tagged data would permit event building to be performed off chip if desired, RD53C builds events on-chip, such that a full event is output before sending any data for the next event. The characteristics of the physical data output ports are described in Sec. 10.1. The transmission protocol used is a subset of the Aurora 64b66b protocol [3], as detailed in Sec. 10.2. This provides industry standard frame alignment, DC balance and multi-lane serial transmission suitable for high speed data, but does not define the data content. The Aurora protocol can be thought of as a "wrapper" placed around the RD53C data. Before the Aurora wrapper, the hit data are packaged in *streams*, not fixed frames. A stream is a self-contained, variable length data container beginning with a tag (8 bits) and followed by a mix of hit data and possibly other tags (called internal tags, which are 11 bits). Streams and their contents are described in Sec. 10.3 to 10.9. This variable length format is approximately 25% more efficient (fewer bits per hit) than the fixed frame format previously used in RD53A.

There are two encoding modes: single chip and multi-chip. Multi-chip encoding must be used when performing data aggregation. The encoding description in Sec. 10.4 is given for single chip mode, and the effect of multi-chip mode is described in Sec. 10.7. The use of multi-chip mode for data aggregation is described in Sec. 11.

The output is highly configurable and must be correctly set up to perform as required. The basic configuration for single chip operation was described in Sec. 3. Control of event size and data filtering options are covered in Sec. 10.8. Use of pre-emphasis for operation with lossy cables is included in Sec. 10.1. Use of test modes, for example for bit error rate studies, is covered in Sec.13. Technical details of clock and data recovery and serialization are given in Sec.14.

10.1 Data Output Drivers

RD53B contains four current mode logic (CML) differential output drivers (Fig. 56) with programmable pre-emphasis. Between 1 and 4 of these drivers will actually be used depending on the Aurora configuration (Sec. 10.2). Each driver is fed by a dedicated serializer circuit that produces the high speed bitstream. The default bitrate is 1.28 Gbps, but it can be reduced in factors of two down to 160 Mbps (and will be the same for all drivers). Serializer details are given in Sec. **??**.

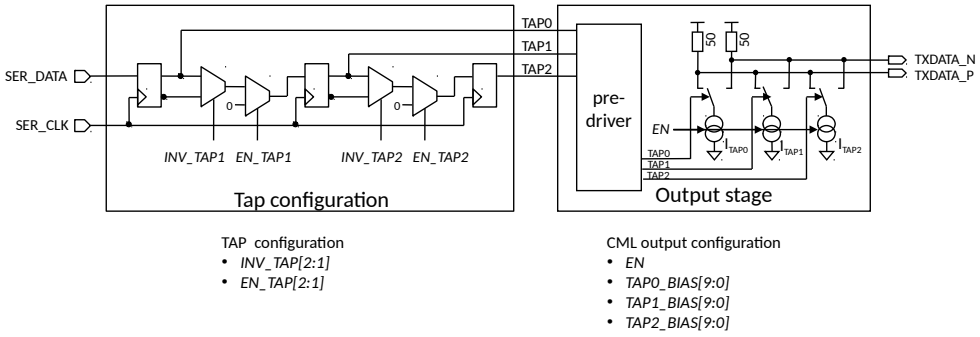


Figure 56: Detailed CML driver functional block diagram including TAP circuit

These differential drivers use back-termination with a 50 Ω pull-up resistor to VDD_CML on each wire to minimize back-reflections and thus improve the signal integrity with non-ideal transmission lines. Each driver can provide pre-emphasis via three current mode switches in parallel (so-called TAPs), which can be programmed to compensate for the high frequency damping of lossy transmission lines using the En. TAP 2,1 and Inv. TAP 2,1 fields of configuration register CML_CONFIG (Table 22). The maximum current for each TAP is approximately 14 mA and the LSB is approximately 14 μA. The TAP configuration (Fig. 56, left) controls the type of pre-emphasis to be used:

- Single TAP: no pre-emphasis
- 2-TAP: programmable overshoot during transitions
- 3-TAP pre-main-post: programmable under-shoot followed by a programmable overshoot during transitions
- 3-TAP main-post1-post2: two levels of overshoot after the transition

The duration of the over/undershoot pulse is fixed by the SER_CLK period, while the amplitude of the output levels can be programmed via the three configuration registers: DAC_CML_BIAS_0 to DAC_CML_BIAS_2. Table 13 shows the configuration settings for the pre-emphasis modes with recommended bias settings. Note that the inversion of TAP 1 is propagated to TAP 2 (Fig. 56).

Pre-emphasis	**Inv. TAP[2:1]**	**En. TAP[2:1]**	**TAP 0 Bias**	**TAP 1 Bias**	**TAP 2 Bias**
off	xx	0	700	0	0
2-TAP (main-post)	x1	1	500	200	0
3-TAP (main-post1-post2)	1	11	500	100	100
3-TAP (pre-main-post)	11	11	100	400	200

Table 13: Configuration settings for the pre-emphasis modes with recommended bias settings. Bias values are register settings (decimal).

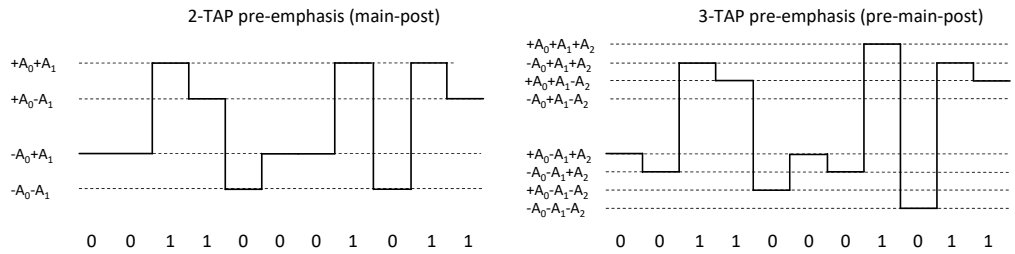


Figure 57: Output waveform with active pre-emphasis in 2-TAP (left) and 3-TAP pre-main-post mode (right). The amplitudes A0, A1, and A2 are controlled by the bias settings of Table 13.

The effect of the pre-emphasis is shown in Fig. 57. In 2-TAP mode, a programmable overshoot is added to every transition to compensate for the high frequency attenuation of the transmission line. The pre-main-post 3-TAP mode adds an additional undershoot in front of each transition which would compensate a higher order low-pass filter transfer function. Tests have shown that 2-TAP pre-emphasis mode gives the best results with the ATLAS prototype cables (6 m long custom twinax cables). Using the bias settings given in the table the 2-TAP pre-emphasis achieves a boost

of 10 dB at 640 MHz (1.28 Gbps). Note that the maximum output amplitude is not limited by the pre-emphasis mode but limited to 700 mV full swing by the saturation voltage of the NMOS current sinks. Thus, the same maximum amplitude is reached with the bias settings shown with pre-emphasis off, as with 2-TAP mode.

10.2 Aurora and RD53C Data

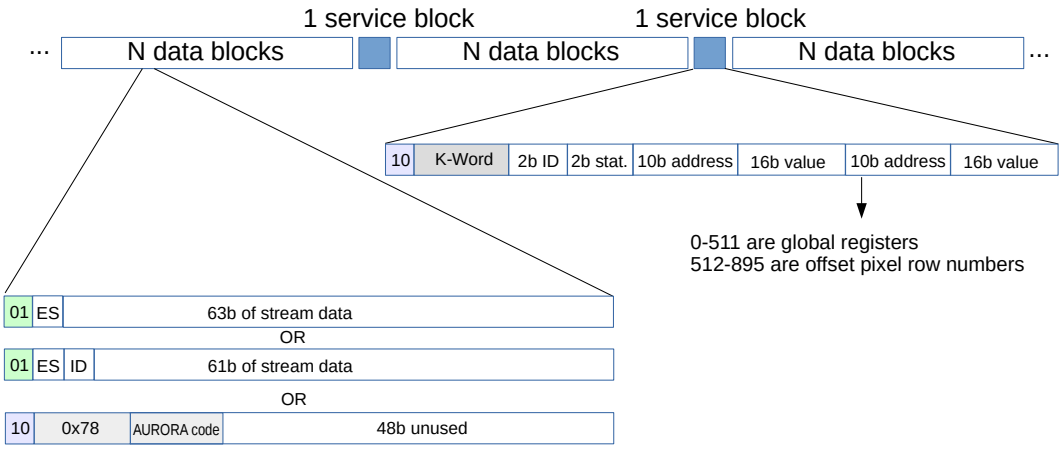


Figure 58: Schematic diagram of output data highest level format, consisting of *N* data or idle blocks followed by one RD53 service block. Each block consists of an Aurora 2-bit header that can only be 01 or 10, plus 64 scrambled bits. The diagram shows the content of the 64 bits before scrambling. The gray shaded 8-bit fields with values given in hex have a meaning defined in the Aurora protocol. The possible Aurora K-Block values are given in Table 14. ES stands for End Stream bit and ID for the two least significant bits of the chip ID.

Aurora K-Word code (hex)	**Meaning**
0xB4	both register fields are of type AutoRead
0x55	first frame is AutoRead, second is from a read register command
0x99	first is from a read register command, second frame is AutoRead
0xD2	both register fields are from read register commands
0xCC	Indicates an error. Fields are meaningless

Table 14: Meaning of Aurora K-Word code (zz) in the periodic service blocks. This table is a companion to Fig. 58

At the highest level, the RD53C output is encoded with a subset of the Aurora 64b66b protocol [3] (see App. A). RD53C implements a simplex channel configuration over 1 to 4 lanes using the *Strict Alignment* feature of the protocol. Pixel hit data are sent in one single infinite length Au-

rora *Data Frame* (binary 01 header), while service data (such as configuration register readback) are sent using Aurora *User K-Blocks* (binary 10 header).

GTX Out	DataMergingMux bits
GTX0	[1:0]
GTX1	[3:2]
GTX2	[5:4]
GTX3	[7:6]

Value (dec)	Internal lane selected
0	0
1	1
2	2
3	3

Table 15: Output switch matrix configuration using register DataMergingMux of Table 22. The right side table indicates which GTX output is connected to the internal Aurora lane based on the programmed 2-bit value.

The ability to configure 1 to 4 output lanes can accommodate different wiring configurations in actual chip usage. It is possible to route any of these four lanes to any of the available output GTX channels of Sec. 10.1. This routing is done configuring the output switch matrix according to table 15. The default value is to route lane 0 to GTX0, lane 1 to GTX1, lane 2 to GTX2 and lane 3 to GTX3.

Aurora blocks consist of 66-bits (Fig. 58). Each block has a 2-bit sync header (01=Aurora Data type, 10=Aurora K-Block or K-Words), followed by 64 scrambled bits. Because the header is not scrambled, it permits frame alignment of the received data. Frame alignment identifies where each 66 bit block starts.

RD53C takes advantage of the differentiation provided by Aurora between Data and K-Blocks to implement two independent output "channels", hit data and service data, as depicted in Fig. 58. These two channels are effectively time-multiplexed onto the serial output. The Aurora encoded output basic unit (which repeats forever) consists of N_D Data or Idle blocks plus one service User K-Block, where N_D is programmable (range 1-256) and has default value 50 (ServiceDataConf register in Table 22). A fraction $1/(N_D+1)$ of the output bandwidth is thus permanently reserved for service information and unavailable for hit data. Conversely, $N_D/(N_D+1)$ is reserved for hit data and cannot be used for service information. If there are no hit data this fraction of the bandwidth will have Aurora idle blocks.

Service blocks will not be sent except in their allocated turn every N data or idle blocks. The interval N is used on every lane regardless of how many lanes are active. For example, with the default N_D=50, 2% of the output bandwidth is permanently unavailable for hit data (in addition to the 3% consumed by the 2-bit 64b/66b header). At 4×1.28 Gbps output bandwidth this 2% is sufficient for the maximum possible register readback of 64 Mbps, since 2% of 5 Gbps = 100 Mbps (See Sec. 8.8). In the service blocks, an 8-bit code follows the sync header, as specified by the 64b/66b protocol, leaving 56 bits available for user information These 56 bits are allocated as a 2-bit chip ID plus two 26 bit registers (10-bit extended address plus 16-bit value = 26 bits) plus 2 status bits, specified in Table 16.

ID[2 bits] 2x([e-address (10 bits)] [value (16 bits)]) [status (2 bits)]

Because of the chip ID, the service block is always compatible with multi-chip mode. The 10-bit extended addresses (e-address) in the service block are: MSB=0, followed by the 9-bit global

register address, or MSB=1, followed by the 9-bit pixel row address in case of reading global register 0 (the pixel configuration portal register). The separation of the output into two time-multiplexed channels guarantees a certain bandwidth for both data and register information without the need for a complex priority arbitration containing safeguards against all possible pathologies.

The periodic service block coming out every N_D data frames is filled automatically, even without there having been a read register command. The possible Aurora K-Words in Fig. 58 are given in Table 14. The two 16-bit registers are denoted A*i* and B*i*, where *i* is the lane number (0 to 3). The automatic filling of the A*i* and B*i* registers is controlled by eight configuration registers Auto-A*i* and Auto-B*i*, which have default values, but which the user is free to change. The auto-fill register addresses are specific to each lane. Thus if only lane 0 is used then only Auto-A0 and Auto-B0 are functional. RdReg commands will queue the registers specified by the command for output on lane 0 only, with priority over auto fill. Lanes 1 to 3 are unaffected by the RdReg command and only output their assigned auto-fill registers. If only one RdReg command has been received, then the A0 register will be auto-filled while the B0 register will contain the requested register. If more than one was received then both registers will be requested registers and auto-fill will wait. If read register commands are sent too fast for the reserved output bandwidth, the FIFO holding pending read registers may fill up, and any read register commands received while the FIFO is full will be ignored. All service block FIFOs have a depth of 16. The readout of registers staged in the FIFO may also happen out of order.

Status Code (decimal)	**Meaning**
0	Ready
1	There has been an error since the last register frame
2	There has been a warning since the last register frame
3	Both 1 and 2

Table 16: Meaning of 2-bit status code

10.3 Aurora and streams

The Aurora protocol transmits data in fixed length blocks with 64 scrambled bits preceded by a 2-bit header. The RD53C encoding does not use fixed length words, but variable length "streams". To fit such variable length streams into fixed length blocks, the first of the 64 bits in an Aurora block, before scrambling, is used as an "End Stream bit" (ES), leaving only 63 bits for actual data. If ES=1, this indicates that the final 63 bits of a stream follow, and the next Aurora block will be the beginning of a new stream. If ES=0, this indicates that the stream continues in the next Aurora block (ie the stream has more than 63 bits to go).

Fig. 59 shows a continuous bit stream as would be seen after Aurora decoding. The position of the ES bits in this bitstream are known (red and blue), thanks to Aurora having taken care of frame alignment. The figure also shows three RD53C streams, which are self-contained, variable length data packets. The last Aurora block of each stream is flagged by ES=1 (red), while ES=0 (blue) only appears in streams that span more than one Aurora block (ie are more than 63 bits long). Note the second stream fits in only one block.

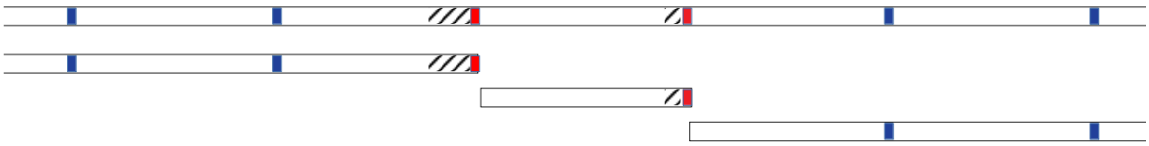


Figure 59: Continuous bitstream after Aurora decoding (top) showing the ES bit positions (which correspond to Aurora block boundaries). ES=1 bits are shown in red, ES=0 in blue, and orphan bits are shown hatched. The three contained streams are shown below, each with its ES=1 bit and orphan bits.

Streams only contain hit and exception data. Configuration readback and monitoring data are not included in streams, but are sent in the periodically inserted Aurora service blocks.

A Stream contains N_E events, where N_E is programmable from 1 to 64. For short or empty events (as will occur in outer layers), single event streams will be inefficient, because the so-called orphan bits at the end of a stream (hatched in Fig. 59) are wasted. For long events (as in the inner layer) single event streams only waste a few percent of bandwidth on orphan bits. The default setting is $N_E = 16$ (in register DataConcentratorConf). Note that single event streams are obtained by programming $N_E = 0$. Even when N_E is programmed >0 single event streams will still occur if the trigger rate is low, as a stream must end when there is no more data to be sent, regardless of N_E.

A new stream always begins with a tag (8 bits) and is followed by a mix hit data and if $N_E > 0$, other tags (called internal tags, which are 11 bits). A tag is always output for every trigger received, even if the event is empty. The possible tag values are given in Table 11. The hit data are compressed and zero-suppressed, and therefore, variable length (number of bits per hit varies).

In addition to the ES bit, the end of a stream can be recognized by a six or more consecutive zeros, which can be neither a valid ccol address nor a valid internal tag (see Sec. 10.4). Orphan bits are always padded with zero. An End of Stream marker function (EoS) forces there to be at least six zeros by adding an entire 64-bit block of all zeros in case a stream happens to end fewer than six bits from the 64-bit boundary. The EoS Marker function can be disabled by zeroing the "EoS marker" configuration bit of register DataConcentratorConf (Table 22).

10.4 Hit data encoding

Within a stream, hit data encoding uses a hierarchical address of core-column (ccol), quarter-core row (qrow) within that column, and 2 pixel x 8 pixel quarter-core hit map, compressed or not as explained later (hit map compression can be disabled in CoreColEncoderConf configuration register). Following the quarter-core hit map are the ToT values for all hit pixels in the quarter-core (which can suppressed by setting bit "Drop ToT" of CoreColEncoderConf register in Table 22. Suppressing ToT will reduce data volume by about 30% at small radius). The order of the ToT values is top row first, from left to right, and bottom row second, from left to right (note that the row number increases from top down). The qrow address field begins with two flag bits called *islast* and *isneighbor*. The islast bit is set if this is the last qrow address in the ccol and zero otherwise, while the isneighbor bit is set if the previous address was qrow-1 and zero otherwise. When isneighbor is set, the qrow address is omitted, as it is known to be the previous address+1. (This is a form of

Huffman coding: since the most frequently occurring qrow address is the previous address+1, due to the clustered nature of hits, a single bit is used to encode this address, while for all other cases a 0 followed by the full qrow address is used.)

Fig. 60 shows the bit content of various hypothetical short streams, without showing Aurora block boundaries. Each of these streams could span one or more Aurora blocks and the ES bits are not shown. These examples illustrate the encoding hierarchy, where the different fields appear depending on the data content, and the functioning of the *islast* and *isneighbor* bits. Placing all ToT's in one block after the quarter-core map makes it simpler to drop ToT, should that be needed, but the default encoding contains the 4-bit ToT values.

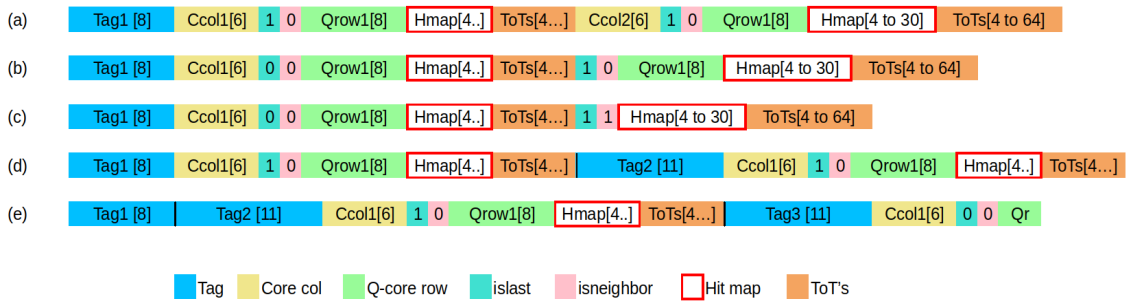


Figure 60: Examples of encoded stream data with no Aurora block boundaries shown and corresponding ES bits suppressed: (a) one hit quarter-core each in two ccols (note last hit bit is set for both), (b) two separated quarter-cores hit in same ccol (last hit set only for second), (c) two neighbor quarter-cores hit in same ccol, (d) one hit quarter-core each in two different events, (e) an empty event followed by an event with one hit quarter-core, followed by another event. A color key to the field types is shown at the bottom. The number of bits in each field is shown in square brackets.

The ccol address is not compressed. The allowed range is 1-55. The value 0 is reserved for the end of stream marker mentioned earlier. Since all valid ccol values are < 56 (binary 111000), an address 111xxx is interpreted as the first bits of an internal tag instead of a ccol. The full internal tag is thus 111xxx xxxxx (see Fig.60d,e). The qrow address begins with the two flag bits islast and isneighbor as explained before. There is only compression in the case of isneighbor=1, which is significant, as this condition is common for clustered hits.

Typically all the data for one ccol will appear together, followed by all the data for another ccol, and so on. But this is not a rule of the encoding. Occasionally, depending on the number of hits and the timing of their extraction from a core column, it may happen that only some of the data for ccol_i appears and is followed by ccol_j, after which more data for ccol_i appears. This is perfectly valid and has important implication for the DAQ. The DAQ must store separately for every ccol the latest qrow value processed (latest qrow must be an array indexed by ccol, not a single variable). This way the DAQ will know where to continue when a ccol value appears more than once. Normally *isneighbor* will be zero for the first qrow address of a ccol, but when that ccol appears a second time in an event readout, it is possible for *isneighbor* of the first qrow to be 1, since the qrow addresses are simply continuing from the first installment of that ccol's readout.

10.5 Stream construction and efficiency

The stream builder (Fig. 55) must decide when to end a stream and start a new one. The builder does not know in advance when a stream will end; the data will determine that. A stream will end when (1) N_E events have been added, or (2) there is no more data to be sent.

For both of the conditions that end a stream, there will typically be a remainder of *orphan* bits between the end of the stream and end of the last Aurora block. These bits could in principle be used for something, but in RD53C they are padded with zeros. The DAQ should ignore orphan bits. For easy identification of orphan fragments the core column addresses start at 1 instead of 0. Thus, 000000 effectively marks the end of a stream, whether the end of stream marker is enabled or not. If the end of stream marker is disabled the number of orphan bits can be fewer than 6, even none, while if end of stream is enabled there will always be 6 or more orphan bits. For example if a stream would have 4 orphan bits (0000) with end of stream marker disabled, a new block would be added, increasing the number of orphan + end of stream marker bits to 67. Fig. 61 shows the bit content of a hypothetical stream extending across two Aurora blocks.

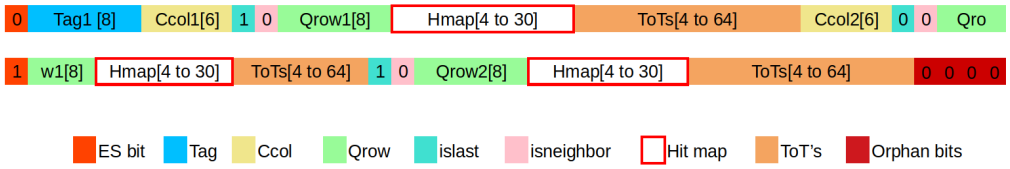


Figure 61: Encoded output for one hit quarter-core in one core column, and two adjacent hit quarter-cores in another core column, spanning two Aurora blocks. The end stream bit (red) is zero for the first Aurora block (top) and one for the second, indicating that the stream ends within the second Aurora block. Orphan bits set to zero (dark red) at the end of the stream in the second Aurora block.

The fraction of bandwidth wasted on orphan bits (inefficiency) can be easily estimated from the stream length. Taking the stream length as an approximately random variable, the average number of orphan (+ end of stream) bits per stream is 31 (37) if end of stream marker is off (on). This, in order to achieve a small fraction of wasted bandwidth, for example <2%, the average stream length must be >1550 (>1850) bits. So one should program $N_E = 1550/\overline{W_E}$, where W_E is the number of bits per event.

10.6 Hit map construction

A 16-bit hit map of the quarter core indicates which of the 16 ToT values are not 1111. The ToT 1111 means "no hit". If the pixel ToT value was 1111, then the corresponding bit in the hit map is zero and otherwise it is one. The default action is to compress the quarter-core 16-bit hit map to use fewer than 16 bits on average. This compression can be turned off in CoreColEncoderConf configuration register (Sec. 16.2). if compression is off, then the hit map will always be exactly 16 bits. In order to compress the hit map, it is (A) encoded using a binary tree and (B) the resulting code is then reduced with a bit code substitution. This section explains the encoding in an algorithmic way that is easy to understand, but does not reflect how it implemented in the chip.

(A) Binary tree construction This is done recursively in 3 steps (for the 16 pixel quarter-core) as follows

1. divide the quarter-core in top and bottom rows and label each row with 1 if it contains any hits and 0 if it does not. The top row is the first bit and bottom row is the second bit.
2. Divide each row of 8 pixels into a left half (first bit) and right half (second bit). The bit is 1 if any of the 4 pixels are hit and 0 if not.
3. Divide each half-row of 4 pixels into a left pixel pair (first bit) and right pixel pair (second bit). The bit is 1 if the pixel pair has a hit and 0 if not.

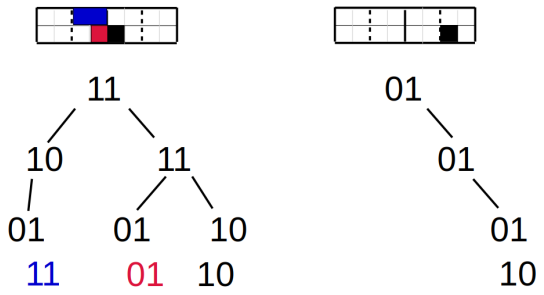


Figure 62: Depiction of binary trees for two example quarter-core maps. The bottom tier of the trees consists of 2-pixel hit maps.

After these 3 steps one has identified all pixel pairs with at least one hit. The 2-bit map for each hit pixel pair is saved (this a 4^{th} step in the chip implementation). The results of the encoding are: 2 bits for step 1, from 2 to 4 bits for step 2, from 2 to 8 bits for step 3, plus the 2-bit maps of all the hit pairs. A quarter-core hit map with a single hit will have an 8-bit binary tree representation. A quarter-core with exactly 2 hits will have a binary tree with between 8 and 14 bits, etc. The maximum number of bits a compressed hitmap with up to 16 hit pixels can use is 30 (2 + 2×2 + 4×2 + 8 2-bit maps). These numbers will all be potentially further reduced by action B.

One necessary ingredient for constructing a tree is a definition of the core subdivisions at each step, as specified in the steps 1-3 above (top-bottom for step 1 and left-right for steps 2 and 3). The trees for two example hit maps are depicted in Fig. 62. Additionally, one must specify in what order the values from Fig. 62 are to be listed. In RD53B the values are listed as follows: The step 1 result (top line Fig. 62) is listed first. This is followed by one step 2 result (the left one if there are two). Then all the step 3 results for this step 2 are listed (can be one or two), followed by all the 2-bit maps associated with those step 3's (can be one to four). A tree will therefore always begin like that: step 1, step 2, step 3. What follows can be another step 3 if there is one, and then the 2-bit maps. If there is another step 2 it will come after the last map for the first step 2. This is summarized as:

```
s1, s2, s3, [s3], map, [3x[map]], [s2, s3, [s3], map, [3x[map]]]
```

The right side of Fig. 62 only contains the minimal number of elements:

```
s1, s2, s3, map
01  01  01   10
```

while the left side contains additional branches:

```
s1,  s2,  s3,  map,  s2,  s3,  s3,  map,  map
11   10   01    11   11   01   10    01    10
```

(B) Bit code replacement. It should be clear from Fig. 62that the bit code 00 never appears, since only maps with at least one hit are being encoded. As there are only three used 2-bit codes, one of them can be replaced with a 1-bit code. The substitution 01 $\rightarrow$ 0 is made everywhere. This is a minimal case of Huffman coding. The encoded maps for Fig. 62 thus become:

- 11 10 01 11 11 01 10 01 10 $\rightarrow$ 11 10 0 11 11 0 10 0 10 (15 bits instead of 18)
- 01 01 01 10 $\rightarrow$ 0 0 0 10 (5 bits instead of 8)

Note that the choice 01 $\rightarrow$ 0 instead of 10 $\rightarrow$ 0 is arbitrary and makes no difference in the data volume for the given choice of subdivisions, as they are symmetric.

The binary tree encoding is elegant because it has an algorithmic form (as described above). However, in the chip it was implemented with an 8-bit lookup table where each binary value is mapped to its encoded value. Had this implementation choice been known in advance, a Huffman encoding could have been used and would have resulted in slightly higher efficiency.

10.7 Multi-chip encoding

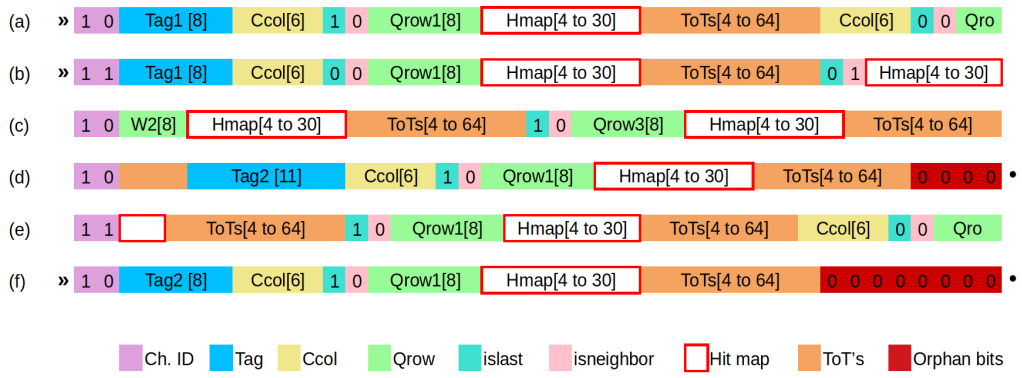


Figure 63: Example of encoded and merged data outputs from two chips with ID LSBs 10 and 11. Six Aurora blocks are shown (a-f), four belonging to chip 01 and two to chip 11 (blocks b and e). The ES bits are not shown, but stream boundaries are shown instead: the start of streams is indicated with the '»' symbol and the end with a dot. Chip 10 has two streams containing 3 events. Event 1 starts in (a) and end in (d). It has two hit Ccols with one hit Qrow in the first and two in the second. Event 2 shares the same stream with event 1. It starts and ends in (d) with only only hit qrow and qcol and is followed by orphan bits. Event 3 starts and ends in (f) in its own one-block stream. Chip 11 has one event with three hit Qrows in the first hit Ccol, the second Qrow being a neighbor of the first. There may also have been Aurora idles or non-data words (such as register readback), which would have been removed or split off by the decoder and are not part of the streams.

Data merging combines data from multiple chips onto a single Aurora output. In this mode each chip still produces streams, but the merged data contains Aurora data blocks from multiple chips. Each 66-bit Aurora block still contains data from only one chip, but blocks from different chips are interleaved. To reconstruct the streams from a given chip, the DAQ must be able to determine which Aurora block belongs to which chip. This is possible thanks two chip ID bits immediately after the ES bit at the start of every Aurora 64 bit block (before scrambling). The presence of these chip ID bits is enabled by default and can be disabled by zeroing bit "Ch. ID" of the DataMerging configuration register in Table 22. Thus, instead of 1/64 overhead from the ES bit, one has 3/64 overhead (ES bit plus 2 ID bits). These two ID bits are the least significant bits of the wire bonded chip ID. All other aspects of the stream encoding remain the same. Fig. 63 shows the bit content of a hypothetical merged data output containing two streams, one from chip ID=10 and another from chip ID=11, extending across multiple Aurora blocks.

Because the stream protocol respects Aurora blocks, the decoder just needs to combine all blocks with the same ID in order to reconstruct the streams from that chip. Multi-chip encoding is the default setting, as the presence of ID bits upon power up will be a nice diagnostic tool even when not using data merging. For maximum data transmission efficiency, single-chip encoding would be selected upon configuring the chip.

10.8 Event size limit and data filtering

Unphysically large events due to exception conditions can cause readout problems and it may be desirable to suppress them. Two levels of truncation are available, applied prior to Aurora encoding, such that unwanted data are discarded as early as possible. The first is at the core-column level, where a maximum number of hits (in multiples of 4) allowed for any single core column can be programmed in the MaxHits[3:0] field of the CoreColEncoderConf configuration register (Table 22).

This feature can be enabled independently for each core column with the EnHitsRemoval_i registers. When enabled, hits in excess of MaxHits will be discarded prior to encoding and the unphysical qrow number 207 will be added with the islast bit set, to tell the DAQ that column truncation has taken place. A hitmap and a single ToT value will be included in qrow number 207 to comply with the encoding format even though they are meaningless. The added compressed hitmap will be binary 0000 and ToT also 0000. This protects against global occupancy extremes, but not against uniform high (but not extreme) occupancy everywhere.

The second truncation mechanism is a readout timeout for each event. If the time elapsed since the readout of an event started reaches a programmed threshold, any further data in that event will be cleared and the event readout will be ended. When this happens, the unphysical qrow number 207 will be added with the islast bit set, to signal to the DAQ that timeout truncation has taken place.

In addition to truncating events too large to be meaningful, it can be desirable to filter out hits known to be backgrounds. A configurable filter is implemented using isolation and/or ToT. However, this feature has a known bug and should not be used. For a description of this function refer to the RD53B manual. The feature is disabled by default.

The order of hit truncation and filtering is as follows. First column hit truncation is performed (if the option is enabled). Then Isolated Hit Removal would be applied on the remaining hits

if enabled (but is should not be enabled), All Hits that survive the filtering process will then be encoded. Hit Removal, based on global timeout, is instead performed while building an event, so after those steps are performed.

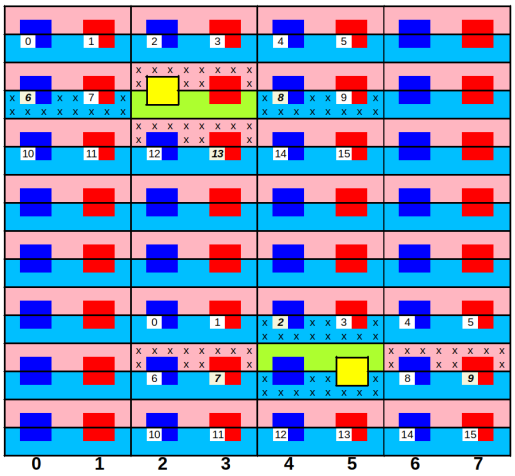


Figure 64: Bump pattern for one core bump bonded to a 25x100 pixel sensor. Column numbers are shown along the bottom. The red (blue) square bump pads are connected to the pink (light blue) pixels. Two pixels/bump pads are highlighted in lime/yellow and the 16 pixels identified by the 16-bit neighbor mask for each are numbered. The up,down,left,right neighbors (hatched) of the upper yellow pixel are selected by setting mask bits 6, 8, and 13; while for the lower yellow pixel by setting 2, 7, and 9. The companion pixel in the same column pair (not numbered) is always set as a neighbor regardless of the mask.

10.9 Precision ToT data

A Precision ToT (PToT) block (Sec. 13.7) is present in every core column and generates four times 16 bits of data, one for each HitOr bus in the core column. These data are stored and triggered the same way as normal hit data, except that the timing can vary: because the HitOr can have a relatively long delay, PToT data can be recorded in the different bunch crossing than the regular hits. The 16 PToT bits consist of 5 Time of Arrival (ToA) bits and 11 ToT bits, as explained in Sec. 13.7.

For the purposes of readout, the 16 bits are considered as a set of four 4-bit fragments. Thus there are sixteen 4-bit fragments. In this way the data can be encoded for readout exactly the same way as a pixel quarter core containing 16 pixels, each with its 4-bit ToT. After encoding there will be a hit map (compressed or not), which is not actually mapping hits but simply indicating which of the 4-bit fragments are not 1111 (since 1111 is the ToT code for no-hit), followed by all the non-1111 4-bit fragments. For example, if the 16-4-bit fragments (expressed in Hex) were 0 0 0 0 5 C F 3 0 0 2 F 0 1 9 2, this will result in the hit map (in binary) 1111 1101 1110 1111 followed by the 4-bit codes (in Hex) 0 0 0 0 5 C 3 0 0 2 0 1 9 2. Only the two F's are missing, and two 0's in the hit map indicate where they belonged. For any given event, some of the four HitOrs may not have fired at all, which is the same as a quarter core without any hits and entirely suppressed from the

readout. Within each 16-bit fragment corresponding to one HitOr bus, the bit assignment is shown in Table 17.

Bit position	0 1 2 3	4 5 6 7	8 9 10 11	12 13 14 15
Decoding	PToT [3:0]	PToT [7:4]	PToA [0:0] PToT [10:8]	PToA [4:1]

Table 17: Bit assignments for the 11-bit precision ToT and 5-bit precision ToA data within the 16-bit fragment corresponding to one HitOr bus.

The ToA and ToT functions are independently enabled with the PToA and PToT bits of the ToTConfig configuration register (Table 22). The 9-Bit field "PToT Latency" of ToTConfig defines a dedicated trigger latency for the PToT block, analogous to the Latency value in TriggerConfig register for regular pixel readout. Additionally, each core column has a dedicated enable bit for its PToT block in registers PrecisionToTEnable_0 to _3. Setting the column-specific PToT enable bit to zero will completely disable the module for that core column, regardless of the PToA/T enable bit values. Moreover, ToT information is dropped from the data output format (Drop ToT bit in Table 18), the PToT data will be also suppressed regardless of the above enable bits. Therefore, it is not possible to combine binary readout with precision ToT.

10.10 Format Options

Option	Register	Default	Description	Section
En. Chip ID	68	on	Include Chip ID[1:0] after ES bit	10.7
EoS marker	74	on	000000 w/enable bit	10.5
En. BCID	74	off	Insert BCID[10:0] after Chip ID.	
En. Trig ID	74	off	Insert trigger ID[7:0] after Chip ID.	
Raw map	75	off	Don't compress hit maps. Map is always 16 bits	10.4
Drop ToT	75	off	No ToT values- hit/no hit information only	10.4

Table 18: Optional elements that can be enabled/disabled and affect stream efficiency. The Register column lists the global register number of Table 22 where the relevant enable bit resides.

The stream format described in so far and exemplified in Fig. 60 is designed for maximum lossless efficiency in bandwidth utilization. More information can be added for debugging or for other functionality when such high efficiency is not needed, or conversely when lower bandwidth utilization must be obtained. Table 18 collects the available options. They can be used in any combination.

11. Multi-Chip Data Aggregation

The RD53C has four differential receivers (Sec. 11.1) that allow one chip (called primary) to aggregate serial data from one or more other chips (called secondaries) and merge it with its own output. The receivers are compatible with the differential data outputs (Sec. 10.1) of other chips. Fig. 65 shows the block-level schematic of the data merging path. The data receivers are designed to work at 320 Mbps, and so the secondary chip outputs must be configured to operate at 320 Mbps, instead of 1.28 Gbps. On the other hand, the primary chip output must be configured high enough to carry all inputs plus its own data. So either 640 Mbps for a single secondary input or 1.28 Gbps for multiple inputs. On an experimental basis the data inputs can be configured to operate at 640 Mbps, in which case the primary chip output must be configured for 1.28 Gbps for a single secondary input or two lanes at 1.28 Gbps each for multiple inputs.

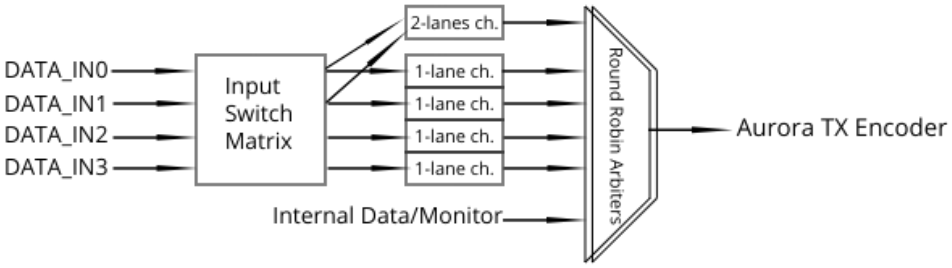


Figure 65: Block level schematic of the data merging path.

11.1 Data Receivers

11.2 Setup and Operation

Lane	DataMergingMux bits
0	[9:8]
1	[11:10]
2	[13:12]
3	[15:14]

Value (dec)	Input selected
0	DATA_IN0
1	DATA_IN1
2	DATA_IN2
3	DATA_IN3

Table 19: Input switch matrix configuration using register DataMergingMux of Table 22. The right side table indicates which input is connected to the lane based on the programmed 2-bit value.

The first block in the data path shown in Fig. 65 is the input selection matrix. With this switch matrix it is possible to select which external serial input is connected to which internal serial lane. This provides flexibility to operate single chip modules in different output modes via configuration, allowing the use of identical modules in different regions of a detector. This configuration is done from the point of view of the internal serial lane: using register `DataMergingMux` according to table 19, each internal lane is fed from a specified data receiver. No sanity check is present to enforce that the selection of inputs is mutually exclusive, or that all connected receivers are used.

Register bit field	Description
[0]	Enable two-lanes Aurora channel using internal lanes 0 and 1
[1]	Enable single-lane Aurora channel using internal lane 0
[2]	Enable single-lane Aurora channel using internal lane 1
[3]	Enable single-lane Aurora channel using internal lane 2
[4]	Enable single-lane Aurora channel using internal lane 3

Table 20: Aurora decoder channels enables using register DataMerging of Table 22. Internal lanes are routed as configured according to Table 19.

All configurations, including unreasonable ones, are possible. For example, one can feed multiple input lanes from a single data receiver, which will lead to duplicate data blocks in the final output.

The internal lanes are then routed to Aurora decoder channels. Lanes 0 and 1 are routed to one two-lane Aurora channel, which interprets these two lanes as a single serial stream from a chip using two output lanes. In parallel, all lanes (0-3) are routed to four single-lane Aurora channels. Each Aurora channel can be separately enabled using the register `DataMerging` of table 20. As before no sanity check is performed and it is possible to enable an unreasonable combination of Aurora channels. Normally either the two-lane channel, or up to three single-lane channels will be enabled- never both. The enable bit provides clock to the channel so that it can function.

For example, one can read three chips on a single output link, by configuring one chip as primary (connected to the output link) and two as secondaries, each with their active output connected to one of the primary's data receivers. Assuming these receivers are DATA_IN0 and DATA_IN2, the primary chip would be configured with DataMergingMux[15:8] = binary xx.xx.10.00, which means internal lane 0 is fed from receiver 00, while internal lane 1 is fed from receiver 10 (little-endian for 1). The values for internal lanes 2 and 3 do not matter (xx), because these lanes will be disabled. Register DataMerging[5:2] is set to binary 0011, which enables the single-lane Aurora channels for internal lanes 0 and 1. Note that in this particular configuration the 1.28 Gbps primary output can never be saturated, as it is fed from three 320 Mbps lanes.

In the case of two chip sharing one 1.28 Gbps output link, there will be one primary chip and a single secondary chip with two outputs connected to two primary chip inputs. Assuming these are DATA_IN3 and DATA_IN1, The primary chip must be configured with DataMergingMux[15:8] = binary xx.xx.01.11 and DataMerging[4:0] = binary 00001 (which selects the two-lane Aurora channel). The data receivers must be routed to lanes 0 and 1 as they are the only lanes used by the two-lane channel. Additionally, the order of lanes matters in the two-lane configuration: output lane 0 of the secondary chip must be connected to the internal lane 0 of the primary chip.

The Aurora input channels are decoded and separated into two data buffers: one for hit data and the other for monitoring. This prepares the incoming data to look just like the internal data from the primary chip before Aurora encoding. Incoming idles, channel bonding blocks or any other Aurora protocol blocks are discarded and not buffered.

Every input lane plus the internal data from the primary chip are then routed to two round-robin arbiters, one for data and the other for monitor. The arbiter will select data from the next non-empty input according to the order: single-lane channel 0, 1, 2, 3, two-lanes channel, internal data path. Note that inputs that are not enabled are guaranteed to be empty, since their clock is

gated off and therefore they cannot be filled. The outputs of the arbiters are fed in input to the primary chip's Aurora encoder block, that at this stage doesn't distinguish the origin of the data (from secondary chips or primary) and transmits a single stream as configured, inserting idles and other Aurora markers as needed. Consider for example a three-chip system with one primary and two secondaries. If all chips are full of data, the final output will have a round robin mix of Aurora blocks: secondary0, secondary1, primary, secondary0, secondary1, primary..., with Aurora markers inserted as needed. However, If only the primary chip has data and the secondary chips do not, the final output will contain all Aurora blocks from the primary chip (primary, primary, primary, primary,...), with Aurora idles inserted only when the primary chip runs out of ready data, as opposed to a fixed time-domain division of the output link giving idle, idle, primary, idle, idle, primary... This is an important consideration when building systems. In particular, it makes it possible for the link sharing to work with different data rates on different chips, as the output will always be filled with whatever input has data. For example if the primary chip produces data at 640 Mbps while two secondaries produce data at 320 Mbps each, half of the 1.28 Gbps output bandwidth will be taken by the primary chip and only one quarter by each of the secondaries.

11.3 Data flow, alignment, and idles

12. Sensing and Monitoring Functions

The Monitoring block in RD53C enables digitization and readout of internal parameters, such as the temperature, the total ionizing dose, and voltages or currents from different parts of the chip and even external ones. Monitoring can be performed at any time, including during data-taking. Monitoring data are transmitted via the normal data output links, time-multiplexed with event hit data (Sec. 10.2). The monitoring procedure entails first selecting what to monitor by routing the given signal to the chip's internal ADC (Tables 26 and 27 show all the available signals), then triggering the digitization action so that the ADC digitizes the signal (Sec. 12.2.5), and finally reading out the digitized value via service data blocks using Read_Register commands or the auto-read function (Sec. 10.2). Temperature and radiation sensors that feed the Monitoring block are distributed as shown in Sec. 2 and described in Sec. 12.3 and 12.4.

The Monitoring block is depicted by the Fig. 66 and contains two sub-blocks:

- An analog current multiplexer followed by an analog voltage multiplexer (MUX)
- A 12 bit Analog to Digital Converter (ADC)

The output of the current multiplexer has a dedicated wire bond pad (I_mux pad), which can be measured externally or turned into a voltage through connection of an external resistor to ground, R_{IMUX} (see Sec. 16.5). The voltage at the I_mux pad is then one of the inputs to the voltage MUX. Another dedicated wire bond pad sources a known current defined by a 10-bit DAC (I_NTC DAC) that can be sent to an external device to ground, nominally an NTC for silicon detector temperature measurement, and so is called NTC_pad. The voltage at NTC_pad is another input to the voltage MUX. The voltage MUX output feeds the ADC, and also has its own dedicated wire bond pad (V_mux pad) for optional external measurement and calibration of the ADC.

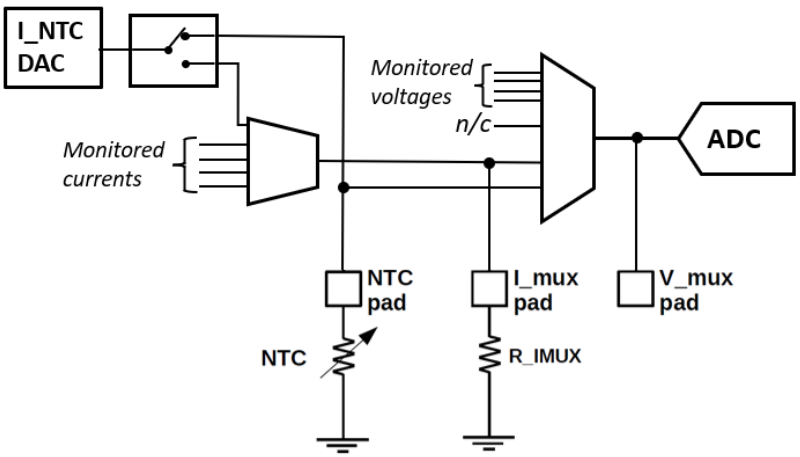


Figure 66: Diagram of monitoring block with current and voltage MUXes feeding the input of the ADC.

12.1 Analog Multiplexer (MUX)

The analog multiplexer has a set of CMOS transmission gates used to connect the analog inputs to a common output. This multiplexer is controlled through selection bits and only one transmission gate is set in the ON state at any given time. Each transmission gate is built with a parallel combination of NMOS and PMOS transistors driven by a complementary gate. When all the selection bits are set to all 1, the multiplexer output is at the high-Z state and the ADC can be calibrated with an external voltage source through the V_mux pad.

12.1.1 Multiplexer Configuration

Global configuration register MonitorConfig is used to enable and select the routing of the multiplexers (See Table 22). The IMUX and VMUX have both a 6 bit selection value. Each value selects a different input, but not all 64 values may used. The list of inputs is given in Tables 26 and 27 of the Reference Sec. 16. Each Mux can be disabled with the output in a high-Z state using the setting corresponding to 63 (all ones).

Examples:

- For monitoring the voltage of the TEMPSENS placed near the analog SLDO (see Sec. 2), channel 14 of the V_mux should be selected (Table 27), so the MonitorConfig configuration register is set to binary 1000000001110 (decimal 4110)
- For Monitoring the NTC current bias, the channel number 9 of the I_mux (Table 26) and the channel number 1 of the V_mux (Table 27) should be selected. The MonitorConfig configuration register is set to binary 1001001000001 (decimal 4673).
- For temperature measurement with an external NTC, the channel number 2 of the V_mux (Table 27) should be selected (after the above has been done, which provided a measurement of the actual current sent to the NTC). The MonitorConfig configuration register is set to binary 1000000000010 (decimal 4098).

12.2 General Purpose ADC

Fig. 67 shows the main circuit elements of the monitoring block:

- The 12 bit General Purpose ADC proper, including the clock divider and the start of conversion signal generation circuit
- The reference voltage selection and buffering
- The input stage analog multiplexer as described earlier

The 12-bit ADC is based on a Successive-Approximation Register (SAR) architecture. It is the most popular architecture for data-acquisition applications, especially when multiple channels require input multiplexing. The circuit takes the chip bunch crossing clock, nominally 40 MHz, and divides it down with a 1024:1 frequency divider to generate the internal clock driving the ADC at 39 kHz. The SAR ADC consists of three main circuits (Fig. 68):

- A 12-bit DAC based on a capacitance network supplied through the reference voltage (Vref_ADC) to generate the voltage scaling

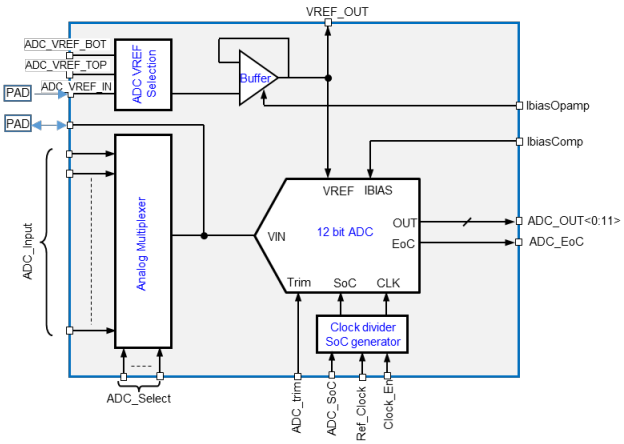


Figure 67: Monitoring block diagram showing ADC.

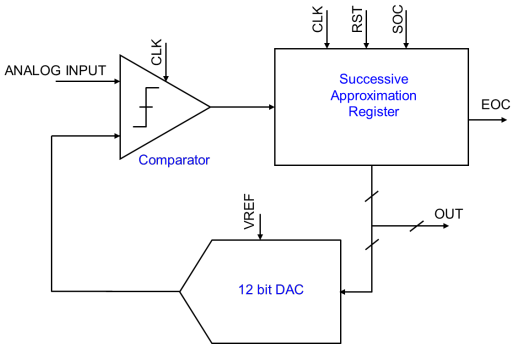


Figure 68: Block diagram of 12-bit SAR ADC.

- A high sensitivity comparator
- A SAR logic block including the frequency divider mentioned above

The ADC digitized output depends on the input and references voltages:

$$\mathrm{ADC_{out}} = A \times \frac{V_{in}}{\mathrm{Vref\_ADC}} + B + \text{nonlinear terms} \tag{12.1}$$

where A is the conversion factor, B is an offset and the nonlinear terms are ideally of order of the LSB or less. Both A and B can be determined from calibration. Prototype measurements with Vref_ADC = 0.9 V have shown a differential nonlinearity of less than ± 1 LSB (200 μV) and an integral nonlinearity less than ± 2 LSB.

12.2.1 12-bit DAC

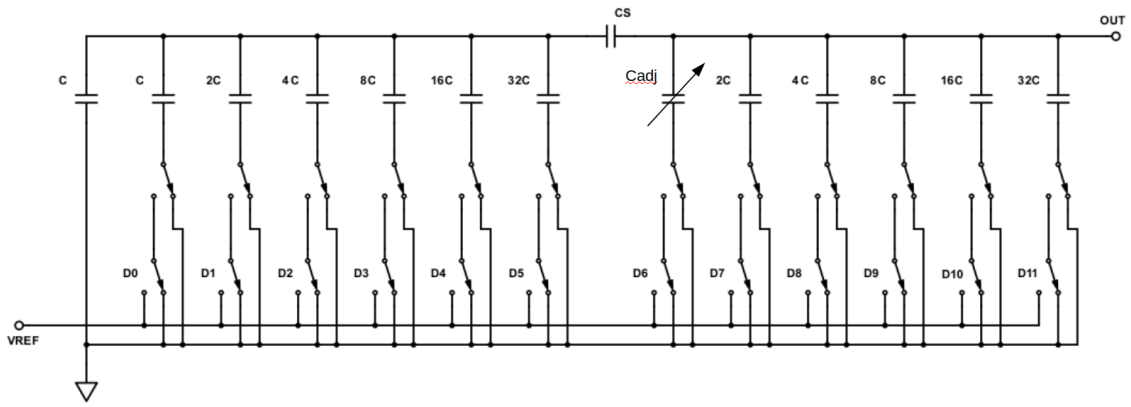


Figure 69: 12-Bit capacitive DAC.

The precision and the linearity of the SAR ADC rely mainly on the capacitive DAC. Two main DAC structures are in general used in SAR ADCs: Binary-weighted and bridge structures.

The bridge structure shown by the Fig. 69 was adopted for this design, as it leads to larger unit capacitance, allowing better element matching and thus higher resolution. However, it suffers nonlinearity caused by mismatch of the bridge capacitance, CS, and by parasitic capacitance in the DAC array. Since this capacitance is insensitive to temperature variation, it can be calibrated to compensate the non-linearity. Six trimming bits are a part of the global register MON_ADC, allowing to adjust the capacitor Cadj between 2C and C/2 in order to compensate the non-linearity from CS and the parasitic capacitance.

The unit integrated capacitance, C, is chosen to keep the mismatch as low as possible and to achieve a very low noise for high accuracy and good linearity. The ratio of output voltage to the reference voltage of the DAC is determined by a capacitance ratio, which makes this stage very tolerant to the radiation damage.

12.2.2 ADC comparator

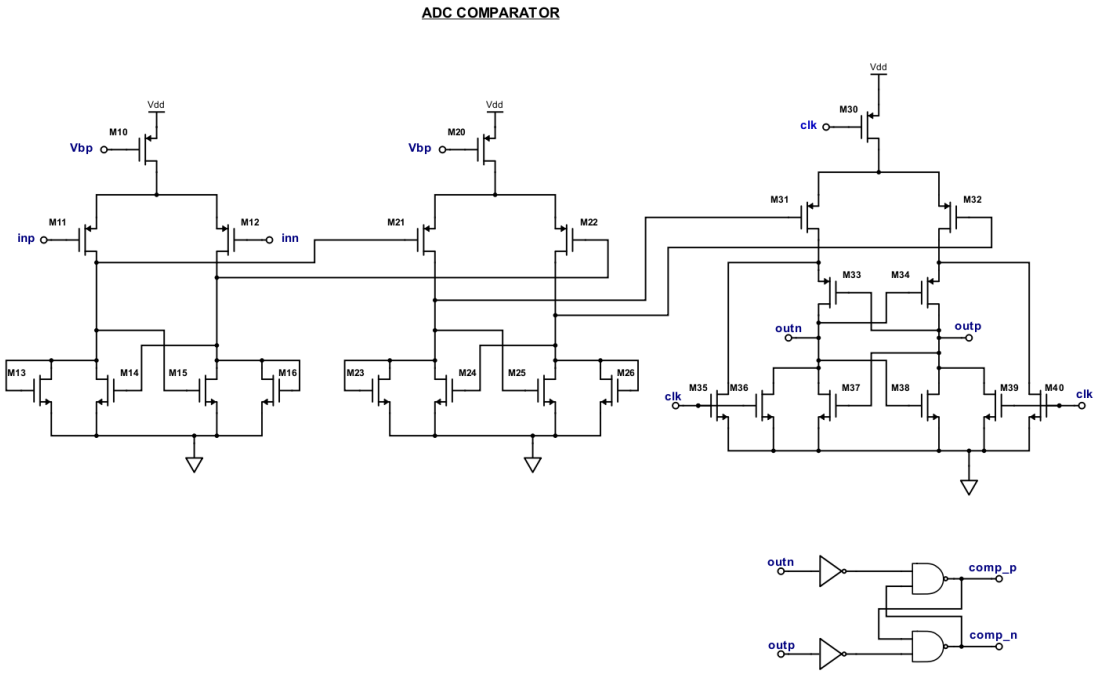


Figure 70: Schematic of ADC comparator.

Fig. 70 shows the 3-stage comparator implemented in the 12-bit ADC design. Two differential operational transconductance amplifiers (OTAs) with diode and current source loads are followed by a dynamic latch comparator.

The first stage input transistors M11-M12 sizes are critical for the linearity and the accuracy. A large transistor area reduces the offset and makes the performance less sensitive to radiation damage, but the gate-source capacitance of a large transistor, which is dependent on the input voltage, would increase non-linearity. A compromise has been found keep low non-linearity with good tolerance to irradiation. There is no switch or reset transistor in the preamplifier stage, and since the voltage gain of the first stage is moderate (10), this keeps the voltage swing at the drain

of the input transistor is kept small. This makes for very small kick-back noise at the input nodes. The dynamic latch comparator has a two phase of operation. When the clock signal is low, the comparator is reset and both the output nodes (outn and outp) are set to 0 V. When clock goes high, a regeneration phase starts and the cross coupled inverter pushes one output to ground and the other to VDD, which goes high and which low depending on the state of the input voltage of M31 and M32.

12.2.3 ADC conversion timing

The timing diagram for the SAR ADC is shown in Fig. 71. Each new conversion cycle is initiated by a Start of Conversion pulse (SOC), which is generated inside the monitoring block. This leads to an ADC start pulse (ADC SOC) with the low frequency ADC_CLK. The ADC_EOC_B flag is asserted when the conversion is completed, indicating that the result can be readout from the ADC data register.

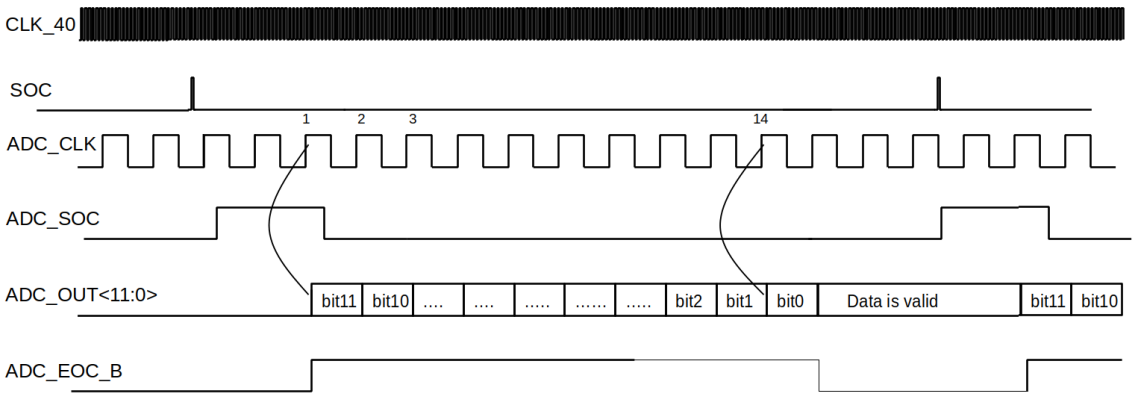


Figure 71: SAR ADC timing diagram.

12.2.4 ADC Configuration

The MON_ADC global register holds the configuration of ADC. The 6 least significant bits hold the capacitor trimming value discussed in Sec. 12.2.1. As these bits primarily compensate for parasitic capacitance and not process, the value is not expected to vary much chip to chip. The 3 most significant bits of the MON_ADC register select one of three reference voltage options. The default value is the Vref_ADC pad voltage, generated by a replica of the main reference current to an external resistor. The 2 other values should be enabled when digitizing the voltages from the resistive temperature sensors (one for each sensor), as shown in Fig. 67.

Since the temperature measurement is a critical function of the monitoring block, the Vref_ADC voltage must be stable in the range -40°C to +60°C. Since another critical function is radiation dose measurement, the Vref_ADC voltage must also be stable vs. radiation dose. This is achieved by using a replica of the main reference current, designed to have such stability, together with an external resistor with negligible temperature coefficient. The external resistance value is chosen to

set the reference voltage around 850 mV. This determines the range and therefore the LSB of the 12-bit ADC.

12.2.5 ADC Control Sequence

The monitoring block is disabled most of the time (MonitorConfig[12] = 0, the default value). The command sequence required to make one conversion is as follows:

- Configure the ADC (see Sec. 12.2.4),
- Write the the MonitorConfig register: set MonitorConfig[12] to 1 to enable monitoring and set MonitorConfig[11:0] to select the channel to be monitored,
- Prepare to start the conversion: write the GlobalPulseConf Register to send the ADC StartOf-Conversion pulse,
- Wait long enough for ADC_EOC_B to go low. For debugging purposes, the ADC_EOC_B signal can be seen on the general purpose CMOS or LVDS outputs.
- Read the ADC result register MonitoringDataADC.
- Disable the monitoring, if no other conversion is to be done.

The conversion time is 14 times the ADC clock period: 358.4 μs. The same conversion result will be read-back by every MonitoringDataADC read command until a new conversion is carried out. The read command itself does not trigger a new conversion (whatever the result was of the most recent conversion will be read back over and over). In case the auto-read register is configured to be the ADC value in order to automatically monitor some value vs. time, one will need to trigger a new conversion periodically in order to be able to see any time variation.

12.3 Transistor-based Temperature and Radiation Sensors

Sensors can be made with any device that exhibits a reproducible temperature dependence. This includes resistors, MOS transistors, diodes, bipolar transistors, delay lines or delay of the logic gates. Most of the commercial temperature sensors are based on the bipolar transistors because the difference in base-emitter voltage is very reproducible as function of temperature. (The bulk CMOS process of RD53B provides parasitic Bipolar Junction Transistors BJTs.) However, these BJTs have been found to be very sensitive to both ionizing dose and displacement damage. Therefore, in RD53B BJTs are used as radiation sensors and diode-connected CMOS transistors are used as temperature sensors. Both cases use the same design of a single device with two switchable bias currents (Sec. 12.3.2) as shown in Fig. 72. This eliminates mismatch on the sensor itself assuming that the temperature and radiation dose are slowly varying.

The sensors have a voltage output that is an input to the VMUX (Sec. 12.1) so they can be digitized by the ADC (Sec. 12.2). The location of the sensors can be found in Sec. 2. Sec. 12.3.1 gives the theory of operation of the sensors.

12.3.1 Transistor Sensor Theory

The voltage V_D across a diode shows a Complementary-To-Absolute Temperature (CTAT) variation. If two biases are applied, I_{bias} and $R \times I_{bias}$, the voltage difference will be given by:

$$\Delta V_D = V_D(R \times I_{bias}) - V_D(I_{bias}) = N_f \times \frac{k_B T}{q} \times ln(R) \tag{12.2}$$

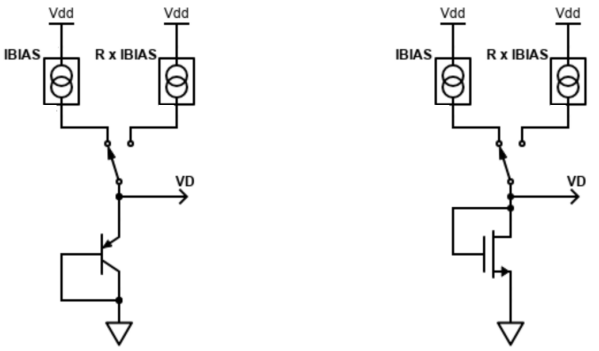


Figure 72: Diagram of BJT radiation sensor (left) and MOS temperature sensor (right) with switchable biases.

where N_f is an ideality factor (1 for an ideal diode), k_B is Boltzmann's constant, T is absolute temperature, and q the fundamental charge. The difference ΔV_D is a Proportional-To-Absolute-Temperature (PTAT). Eq. 12.2 can be rewritten to extract absolute temperature from measured ΔV_D and known R:

$$T = \Delta V_D \times \frac{q}{N_f \times k_B \times ln(R)} \tag{12.3}$$

Formula 12.3 is also valid for a BJT with ideality factor close to 1.0, and for a MOS device biased in the sub-threshold region with ideality factor in the range 1.2 to 1.4. Although the temperature measurement depends logarithmically on R, to have a 1°C precision at 300 K requires a 0.3% precision on ΔV_D and sub-percent precision on R. N_f must also be known precisely, but it is a constant that can be determined from calibration.

12.3.2 Precision Biases

Generation of bias currents with a ratio R is implemented with current mirrors having a ratio R. Despite all the precautions that can be taken at the layout level, it is not possible to achieve an accuracy better than 1% in this ratio. To reduce the error related to the current mirror mismatch, Dynamic Element Matching (DEM) is used. DEM consists of interchanging the unit transistor using a switch array. In the implemented design shown in the Fig. 73: 16 equal current sources generate a 1:15 current ratio, and up to 16 different values of ΔV_D can be measured. The average value is determined offline and the error related to R ratio can be significantly reduced.

12.3.3 Measurement Approaches

Direct For this approach simply measure V_D at the two different bias currents with ratio R and use Eq. 12.3. The digitized ΔV_D is given by $\mathrm{ADC_{out}}(R \times I_{bias})$ - $\mathrm{ADC_{out}}(I_{bias})$. As can be seen from Eq. 12.1, the offset B cancels in this difference, leaving the conversion factor A and Vref_ADC as sources or error. The A nonlinearity is an order 0.2% error. Although Vref_ADC is based on a bandgap circuit, measurements of prototypes showed of order of 3 mV change in the range of -40°C to 40°C. This results in a temperature error of less than one degree, so similar to the A nonlinearity. However, irradiation of prototypes at room temperature showed a shift of -17 mV at 550 Mrad, which is a 2% or 6°C error on temperature.

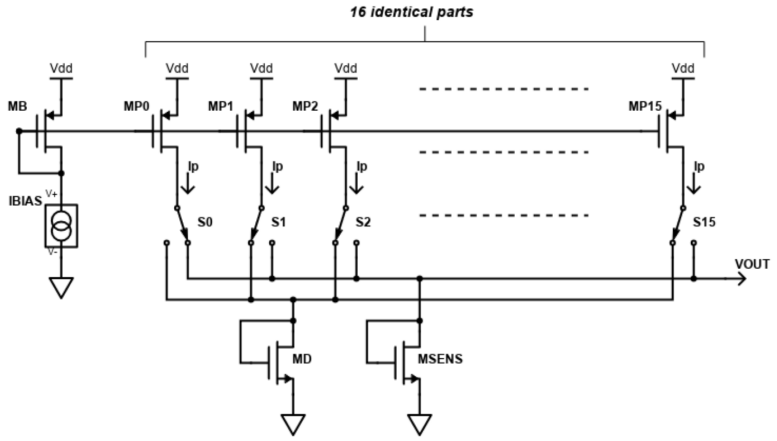


Figure 73: Schematic of Dynamic Element Matching (DEM) circuit to generate precision biases.

Indirect For greater accuracy a self-compensating measurement can be made, using bandgap voltage principle.

12.4 Resistive Temperature Sensors

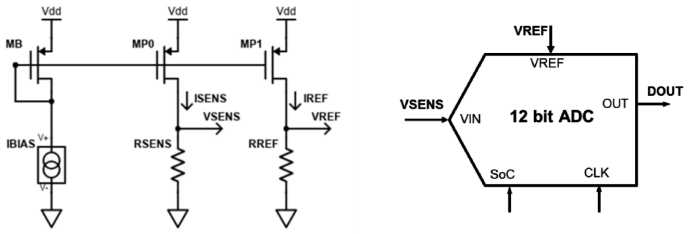


Figure 74: Diagram of polysilicon resistor temperature sensor. To make a differential measurement both an input and a reference voltage are provided to the ADC.

RD53B has two temperature sensors based on polysilicon resistors as indicated in Fig. 4. They are designed to measure the temperature difference between the top and bottom edges of the chip. The small height of these devices allows one to be placed in the very limited available space at the top of the chip. The CMOS process contains various resistor types with different temperature coefficients, allowing sensors to be implemented by comparing the resistance of two resistor types. The simple design for this differential measurement is shown in Fig. 74. Two copies of the same current are passed through two different types of resistor, and the the difference between the voltages across them is measured. Their different temperature coefficients will cause the voltage difference to change in a known way as the temperature changes. The effect is small, so a trick is used to get maximum precision with the ADC. The trick is to use the voltage across one resistor as Vref_ADC and the voltage across the other as the value to be measured. This results in the full 12 bits being available to measure the small difference in the two voltages (see Sec. 12.2.4). The nominal values are 16 K for the resistor connected Vref_ADC and 10 K for

Bits	Field name	Description
MON_SENS_SLDO		
[11]	MON_SENS_SLDOD_EN	336 Enable sensor on digital SLDO
[10-7]	MON_SENS_SLDOD_DEM	Dynamic element matching bits
[6]	MON_SENS_SLDOD_SEL_BIAS	Bias selection switch
[5]	MON_SENS_SLDOA_EN	336 Enable sensor on analog SLDO
[4-1]	MON_SENS_SLDOA_DEM	Dynamic element matching bits
[0]	MON_SENS_SLDOA_SEL_BIAS	Bias selection switch
MON_SENS_ACB		
[5]	MON_SENS_SLDOA_EN	336 Enable sensor on analog SLDO
[4-1]	MON_SENS_SLDOA_DEM	Dynamic element matching bits
[0]	MON_SENS_SLDOA_SEL_BIAS	Bias selection switch

Table 21: Transistor sensor configuration.

the one connected to the ADC input (V_sens). The nominal bias current is 32 μA. With this the condition Vref_ADC>V_sens over the full temperature range for the all process corners.

These resistive sensors can precisely measure temperature changes, but not absolute temperature. Thus, they are ideal to measure the temperature difference across the chip. The ADC value is directly proportional to the relative temperature plus a fixed offset. Measurements on prototypes show good linearity in the range -40°C to 50°C, with a resolution that can reach 5 LSB/°C.

12.5 Sensor Configuration

RD53B has three pairs of active temperature and radiation transistor sensors and two resistance temperature sensors. Their locations were given in Sec. 2. Two 12-bit registers (MON_SENS_SLDO and MON_SENS_ACB) are dedicated to the transistor sensor configuration as shown in Table 21. These control the sensors on the SLDO regulators and chip bottom center, respectively. To perform the measurement:

- Set the sensor enable bit,
- Set the bias switch to 0 (I_{bias}),
- Cycle through all the DEM values and measure the voltage for each value. Average all the measurements,
- Set the bias switch to 1 ($15\times I_{bias}$),
- Cycle through all the DEM values and measure the voltage for each value. Average all the measurements.

The resistor temperature sensors do not have any configuration bits. However, the ADC must be specially configured in differential mode as explained in Sec. 12.2.4.

13. Test and Miscellaneous Functions

13.1 General purpose LVDS and CMOS outputs

RD53C contains four LVDS differential outputs and one CMOS output mainly for testing. During detector operation these outputs enable command link sharing, which means one chip can be used as a repeater for the command serial link towards downstream chips (Sec. 3.1). These outputs also provide early diagnostics (default setting). They can be configured away from default to "spy" on a number of internal signals as detailed in Table 28.

13.2 Bypass mode

Bypass mode allows to control the chip without the internal PLL Clock and Data recovery function. This mode can only be selected by driving a wire bond pad to high. It is available only for expert use to characterize performance of internal blocks.

13.3 Scan Chains

RD53C includes Design For Test (DFT) methodology in the digital flow. This allows structural testing of the bottom of chip logic and also much of the pixel matrix core logic. The General purpose LVDS I/O as well as data merging inputs are used for DFT "scan chain" testing, which is intended to be done at the wafer probing stage. A detailed (technical) description of the DFT functionality is given in App. **??**.

13.4 Hit OR

Within each RD53C core column there are four independent Hit OR nets, each one fed by one quarter of the pixels. Fig. 75 shows graphically how the 64 pixels in one core are grouped into the 4 OR networks. The figure also indicates two possible sensor formats of 50 μm $\times$ 50 μm (50x50) or 25 μm $\times$ 100 μm (25x100) pixels. It can be seen that in the 50x50 case, a given pixel in network 1 has its two up-down neighbors on network 3, and its left-right neighbors on 2 and 4. Conversely, a given 25x100 pixel on network has has its left-right neighbors in network 3 and its up-down neighbors on 2 and 4.

Each net forms the logical OR of all individual pixel outputs that have been enabled by the HitOr mask bit (one bit per pixel). Because the signals travel in an OR network, there will be a different delay depending on which pixel core the signal comes from: the difference between top and bottom of chip is around 6 ns in unirradiated chips at nominal 1.2 V digital voltage. The delay varies linearly with pixel row.

13.5 Heartbeat and test patterns

13.6 Ring Oscillators

RD53C contains a large variety of ring oscillators mainly intended to allow characterization of logic cell radiation tolerance, but which also allow other measurements, such variation of the pixel injection capacitance value. The ring oscillators are located in two banks in the chip bottom: ROSCA and ROSCB, as specified in Sec. 2. Bank A (ROSCA) is a copy of the ring oscillator bank in the RD53A chip, so that a direct comparison with RD53A test results can be made. The two

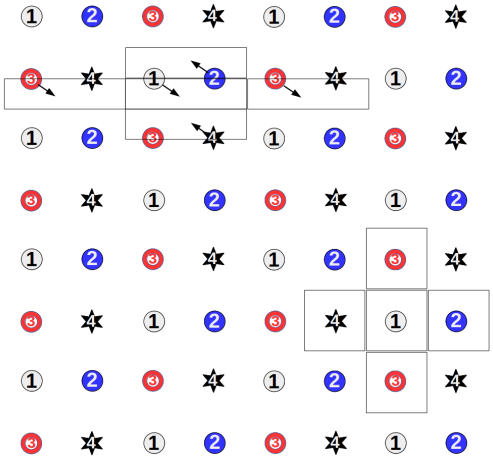


Figure 75: The four Hit Or nets in a 64 pixel core.

banks with their control signals are shown in Fig. 77, while a single oscillator diagram is shown in Fig. 76.

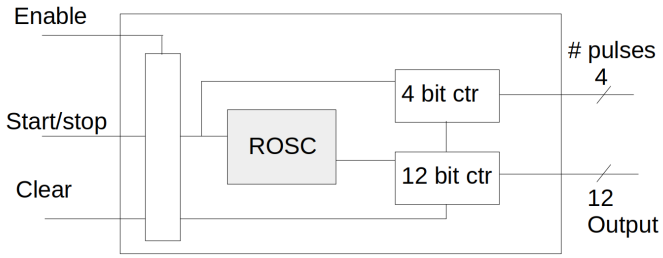


Figure 76: Diagram of a ring oscillator block. Different numbers and types of logic cells are used as specified in Tables 37, 38.

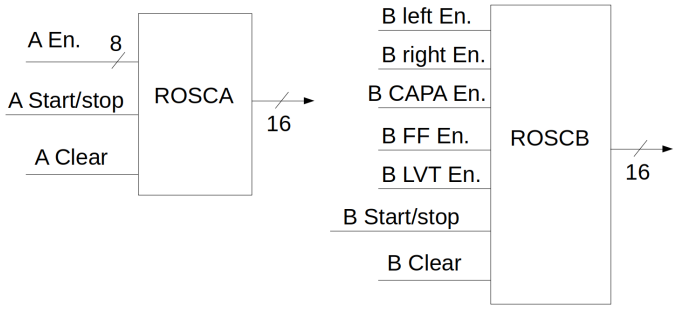


Figure 77: The two banks of ring oscillators with their control signals.

Each ring oscillator is a chain made of different logic cells (Tables 37, 38). The number of cells in each ring was chosen to have approximately the same frequency, as given in the tables. Each oscillator drives a 12-bit counter, while a 4-bit counter is used to count the number of start/stop

pulses received. The counters will count while start/stop pulse is high as long as the Enable bit is high. The start/stop pulses are supplied using the Global Pulse command while Enable and Clear are static configuration bits. The 16-bit counter values from each oscillators are assigned to a read-only global configuration register for readout. There is one register serving the bank A (119: RING_OSC_A_OUT) and one serving bank B (120: RING_OSC_A_OUT).

When Enable is Low the counters hold their values. For bank B only, the clear state also causes the gates to be continually clocked at 40 MHz (not shown in the diagrams). This allows irradiation while the gates are being clocked rather than in a static value. The oscillators within a bank share some control signals and their 16-bit outputs are multiplexed. Bank A contains only 8 oscillators which have individual enable signals and common clear and start/stop. Bank B contains 4 groups of oscillators and each group has its own enable signal, common to all the oscillators in the group (see Table 38), while the clear and start/stop signals are common to the whole bank. The total number of oscillators in bank B is 34. All select and clear bits for both banks are contained in configuration register RingOscConfig (number 117). The A and B start/stop pulses are separate Global Pulse channels (Sec. **??**). Which register is connected to each bank's output is selected with global register RingOscRoute (118), using the oscillator numbers from Tables 37, 38.

13.7 Precision ToT module

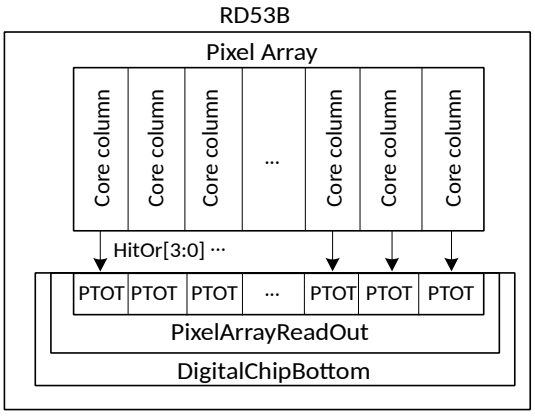


Figure 78: PTOT modules, one per core column.

The Precision ToT (PTOT) module makes measurements on the HitOR signals (13.4) coming out of each core column. There is one PTOT per core column as shown in Fig. 78. Two quantities are measured:

- ToT, just like in the pixels, but with an 11 bit counter counting as 640 MHz rate,
- Time of Arrival (ToA) of the leading edge, as a phase difference from the previous BX clock rising edge to the HitOr leading edge, with a 5 bit counter counting at 640 MHz rate.

These quantities are stored in memories just like the pixel region memories, associated to latency buffers that keep track of time just as in the pixel regions. A diagram the PTOT module

is shown in Fig. 79. The latency buffer depth and logic are the same as in the pixel regions. The readout of the PTOT data is trigger based, exactly as for regular pixels, and the data are included in normal data path as described in Sec. 10.9.

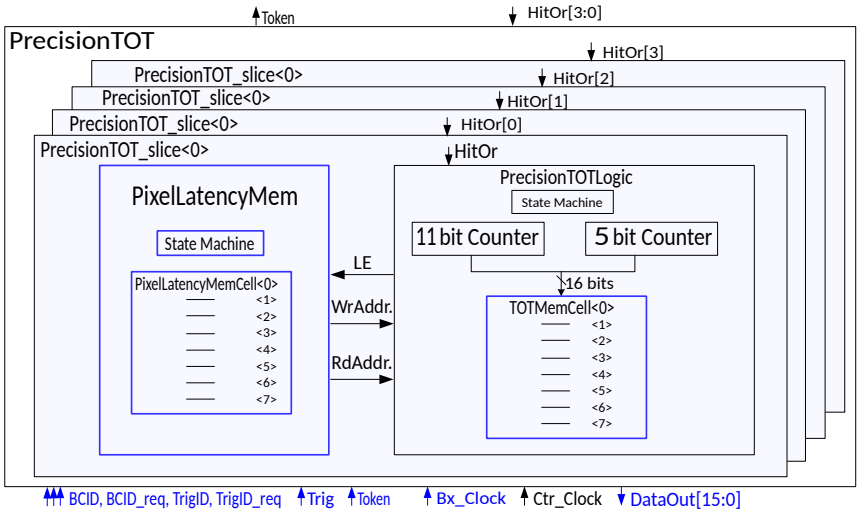


Figure 79: Single PTOT module diagram. The elements outlined in blue are copies (same code) of the pixel region logic.

13.8 Capmeasure circuit

The capacitance measurement circuit (capmeasure, Fig. 80) allows the determination of the as-built front-end injection capacitor C_{inj}. It is integrated into the calibration block in the Analog Chip Bottom. The circuit consists of two sections: the capmeasure and parasitic capmeasure. The capmeasure circuit is connected to a parallel array of 100 capacitors, each identical to the injection capacitors, but connected differently than in the pixel (see Fig. 81) for use as a charge pump rather than a single charge injector. Also, in order to make an array, routing metal is needed, which adds a parasitic capacitance. The total capacitance measured by this circuit is then,

$$C_{meas} = 100(C_{pix} + C_{shld}) + C_{par} \tag{13.1}$$

where C_{pix} is the mutual capacitance between input and output terminals as seen by a pixel, C_{shld} is the added capacitance between input terminal and the shield layer underneath due to the different connection illustrated in Fig. 81, and C_{par} is the parasitic capacitance due to array metal routing.

An identical array but with 50 capacitors having the two main terminals connected together as in the right scheme of Fig. 81 and 50 capacitors left unconnected is therefore also provided in order to measure the shield and parasitic capacitance:

$$C_{meas2} = 50(2xC_{shld}) + C_{par} \tag{13.2}$$

From the two measurements, the injection capacitance can be obtained:

$$C_{pix} = (C_{meas} - C_{meas2})/100 \tag{13.3}$$

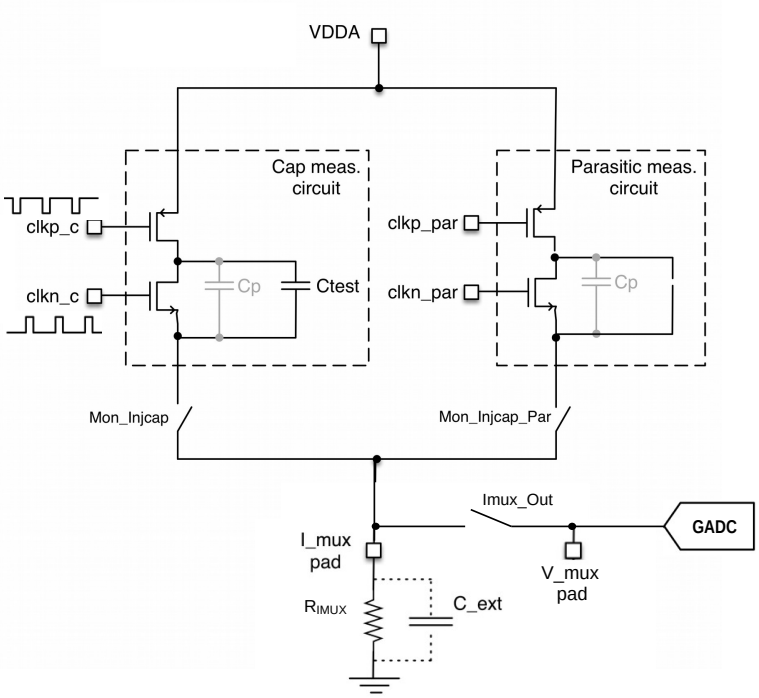


Figure 80: Capacitance measurement circuit diagram. C_{test} is an array of 100 injection capacitors. Control of the switches Mon_Injcap, Mon_Injcap_Par, and Imux_Out, as well as operation of the GADC generic ADC are described in Sec. 12.2.

As for reference, the extracted value from the chip layout is C_{pix} = 8.02 fF.

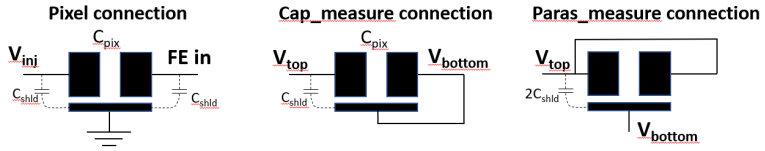


Figure 81: The pixel injected charge is defined by the mutual capacitance between the two main terminals of the injection capacitor (left). However, the capacitor has a third terminal which connects to a shield layer underneath (a poly layer usually connected to GND). Both main terminals exhibit a parasitic capacitance to this node. The capmeasure circuit charges this shield capacitance in addition to the capacitance between the two main terminals (center). In the parasitic measurement circuit, the capacitor has the two main terminals connected together (right)

The capmeasure circuit is based on a charge pump with PMOS and NMOS transistors controlled by non-overlapping clocks. These clocks run at 1/4 of the bunch crossing clock (which is nominally 40 MHz). They are generated by combinatorial logic from the bunch crossing clock and are disabled by default. The En_injcap_meas and En_injcap_par configuration bits (there are actually two control circuits, one for the array of the replica capacitors and one for the empty array to measure C_{par}).

In a simulation of the circuit as shown in Fig. 80 with R_{IMUX} = 5 K and C_ext = 22 nF, the average current flowing in VDDA of the capmeasure circuit is 11 μA for the 10 MHz nominal clock frequency, while the average current flowing in VDDA of the parasitic circuit is 1 μA. This can be sensed by measuring the voltage across the external resistor with the generic ADC. The 22 nF capacitor in parallel will keep the output voltage constant for the GADC measurement (it is possible that the measurement will still be accurate without it- to be checked in bench tests).

The measurements of C_{meas} and C_{par} are made separately. The capmeasure circuit should be reset before making a measurement. This is a achieved by selecting capmeasure in the global pulse routing and issuing a global pulse with a 3 clock width (this is longer than the default width of 1 clock). For C_{meas} (C_{par}), first enable the clock by setting En_injcap_meas = 1 (En_injcap_par = 1) and selecting the IMUX channel Mon_injcap (Mon_injcap_par). One must wait at least 0.5 ms for the output to settle (with the nominal 5 K and 22 nF external components). At this point the voltage at the Imux_Pad can be measured either with an external instrument or the GADC. To use the GADC select input channel Imux_Out and follow the procedure from Sec. 12.2.

In terms of the measurements, the injection capacitance is given by Eq. 13.4,

$$C_{\mathrm{pix}} = \frac{1}{100}\left(\frac{V_{\mathrm{meas}}/R_{\mathrm{IMUX}}}{f \times (VDDA - V_{\mathrm{meas}})} - \frac{V_{\mathrm{meas2}}/R_{\mathrm{IMUX}}}{f \times (VDDA - V_{\mathrm{meas2}})}\right) \tag{13.4}$$

where V_{meas} and V_{meas2} are the measured GADC voltages for the capmeas and parasitic circuits, respectively, R_{IMUX} is the external resistor used to convert IMUX output current into voltage, and f is the charge pump frequency of 10 MHz.

14. Clock Generation and Data Recovery Technical Details

15. Known Issues

This appendix collects information about known bugs or strange features in the design. None of the items in section prevent full operation meeting all requirements. However, the DAQ system must be aware of these issues to properly operate the chip. Boldface items require specific DAQ action and cannot just be ignored.

- Isolated hit removal. Under high hit load fails to remove hits or removes too many hits. This feature should not be used. It is off by default.

- **Suppressed read register**. Read register command may be ignored when a previous read pixel register is being processed. This is a feature of how the commands work. The DAQ must allow time for read pixel commands to finish before reading something else.

- Chip stuck when filter removes all hits. When a core column has triggered hits, the data concentrators that are the first step in event building will be expecting hits from that core column. Therefore, if no hits come out of the column, the chip may become stuck. A clear command or equivalent will restore normal operation. For example, the isolated hit removal function (which should not be used in any case; see above) could remove all hits that were present in the column. A single event upset could also occasionally remove a hit. The vulnerability to this type of SEU has been greatly reduced in RD53C, so it should be very rare.

- Aurora clock compensation CCWAIT. This parameter does not function properly and causes Xilinx Aurora receivers to have errors for the default value of CCWAIT. This has no impact on operation because the clock compensation feature of the Aurora protocol is not used by RD53C. Custom receiver firmware, such a implemented in YARR and required in the experiment DAQ systems will not have any sensitivity to this issue.

- **Aurora channel bonding CBWAIT**. This parameter's default value will cause errors in channel bonding for multilane chip readout (inner layers). Channel bonding is a used part of the protocol. The solution is to write a different value of CBWAIT than the default. The correct value to be determined by systems testing.

- **Clear command reset too short**. The CLEAR command clears the data path and resets the Aurora protocol. However, the reset is only held for one clock cycle, and this is not sufficient to correctly lock timing in different clock domains. This is mainly an issue when data merging is used, and not for single chip operation. The solution is for the DAQ to issue global pulse commands in place of CLEAR commands, with the pulse duration set longer (eg. 8 clock cycles). The same resets as actuated by CLEAR can be selected by global pulse configuration, so the functionality is exactly equivalent, other than the length of the reset pulse.

- **Data merger failure to decode secondary chip data**. For a primary chip to correctly receive the data from a secondary chip it must first find the Aurora frame alignment of the incoming data. The standard method is used of looking for the 2-bit Aurora headers and locking the

phase where all headers are valid. A 64-bit search window was incorrectly implemented for this search, which is too short because the headers happen every 66 bits. This bug means that, given arbitrary frame alignment, the primary chip will fail to lock to the correct secondary chip phase 1/33 of the time. This work-around is to force a non-arbitrary frame alignment by always resetting all chips in a data merging group at the same time, which is trivial because they share the command link. One can either rest all chips in broadcast mode or ensure a deterministic time difference between the reset. Never reset an individual chip in a data merging group. Note that because the primary must always be reset, this will introduce a 20 μs dead time upon reset, while the primary resyncs.

16. Reference Tables (pinouts, configuration, etc.)

16.1 Wire Bonding Pinout

The pinout is provided in a full page figure (next page, no figure number). The 198 pads are shown in two halves just to fit the page. All pads are shown and power and ground pads are color coded while all other pads are shown as open rectangles. Note that every 5^{th} pad is numbered on the actual chip as shown in Fig. 7. Figures 7 and 8 are reproduced here for convenience (note the original figure numbers have been kept).

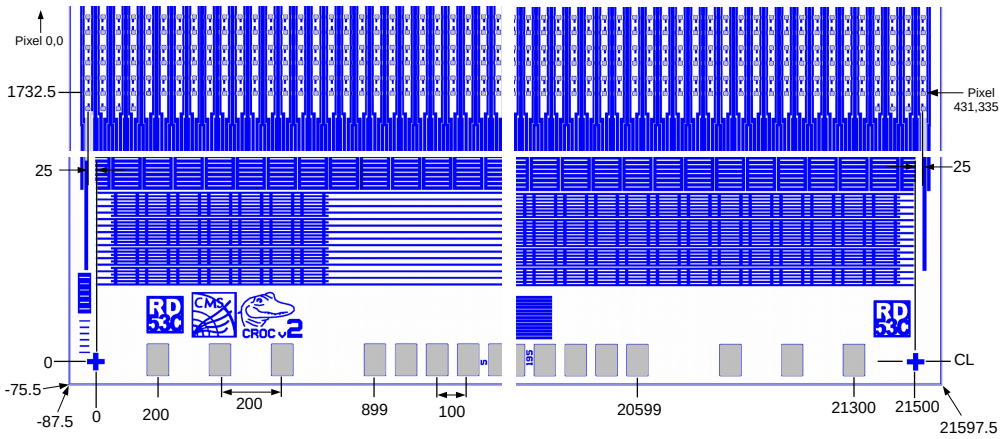


Figure 7: Detail of CMS chip bottom with dimensions (rounded to nearest micron). The pinout follows on the next page, rotated clockwise 90° relative to this figure (pin 1 is at the bottom left of this figure). The location of the first and last pixel bump bonds on the matrix is also indicated. There are 4 bump bond pads below the full matrix on each of left and right sides to contact sensor bias/guard rings.

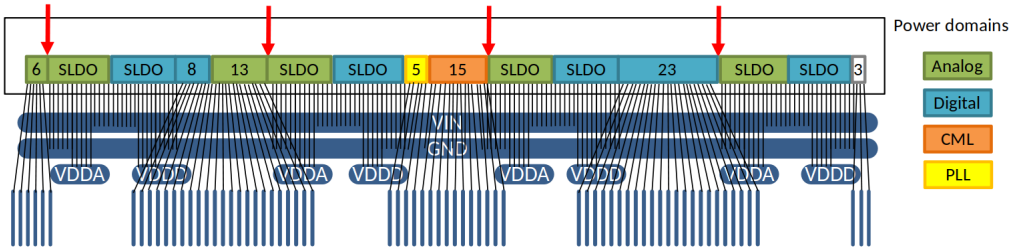
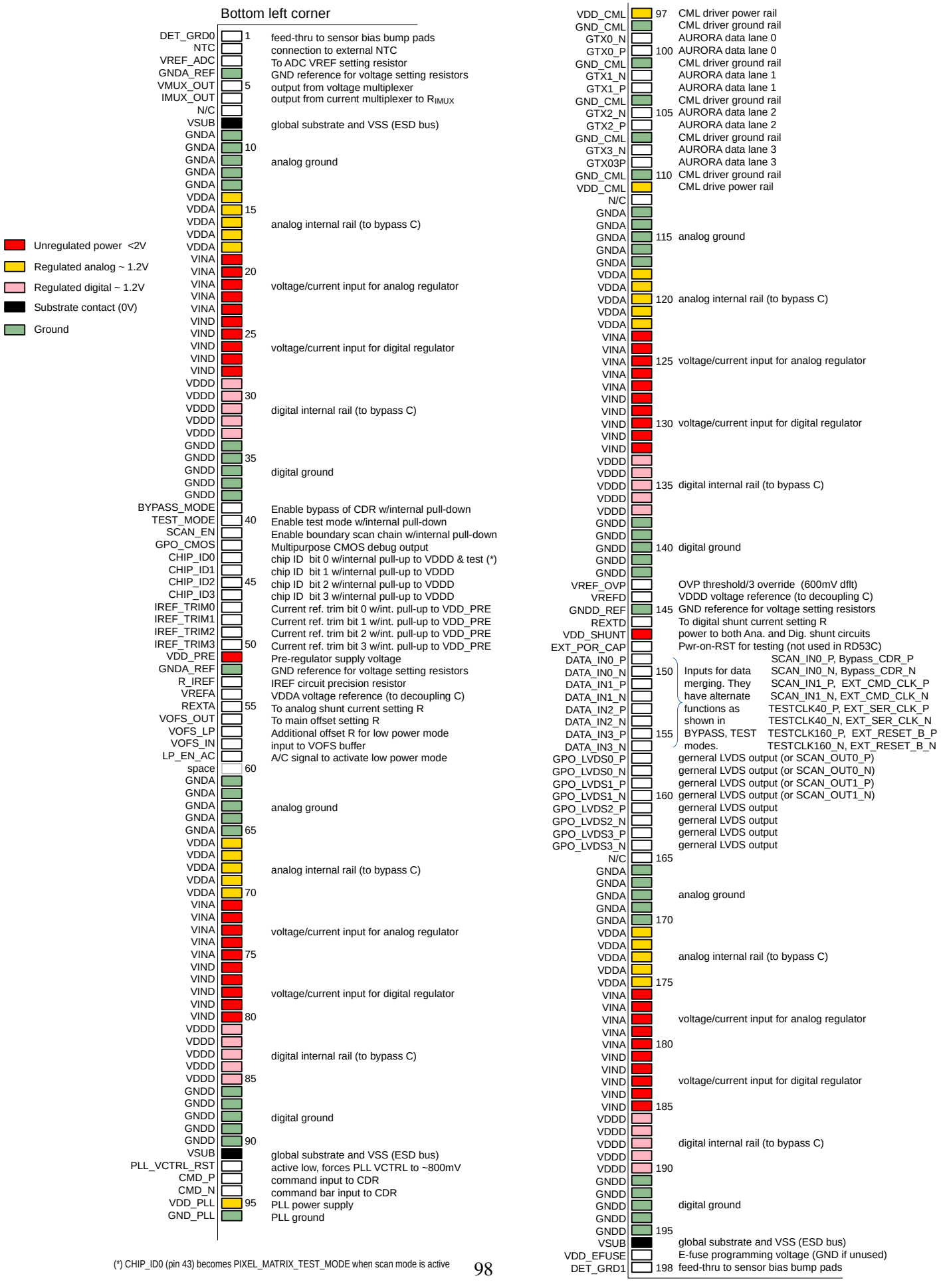


Figure 8: Organization of wire bond pad frame and generic bonding scheme. All wire bonds are shown, including connections for testing (not used on detector modules). the number of fanned-out signal bonds is written in each box, while the power supply bonds run parallel (not fanned out). The red arrows indicate the four unused pads. Full pinout follows on next page.



(*) CHIP_ID0 (pin 43) becomes PIXEL_MATRIX_TEST_MODE when scan mode is active

16.2 Global Configuration

Table 22: Global configuration register main table. (*) indicates further details are given following the table. Section references are given in (). The Size column may list multiple fields by their size in bits. The rightmost value is always the field starting at register bit [0] (i.e. all words are little-endian). Thus, the numbers "1,3,2" in the Size column would indicate a word with 3 fields: [5],[4:2],[1:0]. n/u Stands for not used in ATLAS chip.

Addr.	Name	Size (bits)	Description	CMS Default
* 0	PIX_PORTAL	16	Pixel portal: virtual reg. to access pix config (8.8)	0
1	REGION_COL	8	Pixel column pair connected to pixel portal (8.8)	0
2	REGION_ROW	9	Pixel row connected to pixel portal (8.8)	0
* 3	PIX_MODE	5,1,1,1,1,1	n/u, n/u, Broadcast, Wr_cfg, Auto Row (8.8)	0,1,0,1,0
* 4	PIX_DEFAULT_CONFIG	16	Key 1 of 2: hex 9CE2 to exit pix default config.	0
* 5	PIX_DEFAULT_CONFIG_B	16	Key 2 of 2: hex 631D to exit pix default config.	0
* 6	GCR_DEFAULT_CONFIG	16	Key 1 of 2: hex AC75 to exit glob default config.	0
* 7	GCR_DEFAULT_CONFIG_B	16	Key 2 of 2. hex 538A to exit glob default config.	0
			Analog Front End (5)	
8	DAC_PREAMP_L_DIFF	10	not used in RD53C _CMS	50
9	DAC_PREAMP_R_DIFF	10	not used in RD53C _CMS	50
10	DAC_PREAMP_TL_DIFF	10	not used in RD53C _CMS	50
11	DAC_PREAMP_TR_DIFF	10	not used in RD53C _CMS	50
12	DAC_PREAMP_T_DIFF	10	not used in RD53C _CMS	50
13	DAC_PREAMP_M_DIFF	10	not used in RD53C _CMS	50
14	DAC_PRECOMP_DIFF	10	not used in RD53C _CMS	50
15	DAC_COMP_DIFF	10	not used in RD53C _CMS	50
16	DAC_VFF_DIFF	10	not used in RD53C _CMS	100
17	DAC_TH1_L_DIFF	10	not used in RD53C _CMS	100
18	DAC_TH1_R_DIFF	10	not used in RD53C _CMS	100
19	DAC_TH1_M_DIFF	10	not used in RD53C _CMS	100
20	DAC_TH2_DIFF	10	not used in RD53C _CMS	0
21	DAC_LCC_DIFF	10	not used in RD53C _CMS	100
22	DAC_PREAMP_L_LIN	10	Input transistor bias for left 2 columns	150
23	DAC_PREAMP_R_LIN	10	Input transistor bias for right 2 columns	150
24	DAC_PREAMP_TL_LIN	10	Input transistor bias for top left 2x2	150
25	DAC_PREAMP_TR_LIN	10	Input transistor bias for top right 2x2	150
26	DAC_PREAMP_T_LIN	10	Input transistor bias for top 2 rows	150
27	DAC_PREAMP_M_LIN	10	Input transistor bias for all other pixels	150
28	DAC_FC_LIN	10	Folded cascode branch current	20
29	DAC_KRUM_CURR_LIN	10	Krummenaker feedback current (return to baseline)	70
30	DAC_REF_KRUM_LIN	10	Krummenacher reference voltage	360
31	DAC_COMP_LIN	10	Comparator current	80
32	DAC_COMP_TA_LIN	10	Comparator output stage current	900
33	DAC_GDAC_L_LIN	10	Global threshold for left 2 columns	900
34	DAC_GDAC_R_LIN	10	Global threshold for right 2 columns	900
35	DAC_GDAC_M_LIN	10	Global threshold for all other pixels	900
36	DAC_LDAC_LIN	10	Local threshold current	130
* 37	LEAKAGE_FEEDBACK	1,1	not used in RD53C _CMS	0,0
			Internal Power (4)	
38	VOLTAGE_TRIM	1,1,4,4	Regulator: En. Undershunt A, D; Ana. V trim, Dig. V trim	0,0,8,8

Table 22: Global configuration register main table. (*) indicates further details are given following the table. Section references are given in (). The Size column may list multiple fields by their size in bits. The rightmost value is always the field starting at register bit [0] (i.e. all words are little-endian). Thus, the numbers "1,3,2" in the Size column would indicate a word with 3 fields: [5],[4:2],[1:0]. n/u Stands for not used in ATLAS chip.

	Addr.	Name	Size (bits)	Description	CMS Default
			Pixel Matrix Control		
	39	EnCoreCol_3	6×1	Enable Core Columns 53:48	0,1,0,0,1,0
	40	EnCoreCol_2	16×1	Enable Core Columns 47:32	0,5x(1,0,0)
	41	EnCoreCol_1	16×1	Enable Core Columns 31:16	5x(1,0,0),1
	42	EnCoreCol_0	16×1	Enable Core Columns 15:0	5x(0,0,1),0
	* 43	EnCoreColumnReset_3	6×1	Enable Reset for core cols. 53:48	0
	* 44	EnCoreColumnReset_2	16×1	Enable Reset for core cols. 47:32	0
	* 45	EnCoreColumnReset_1	16×1	Enable Reset for core cols. 31:16	0
	* 46	EnCoreColumnReset_0	16×1	Enable Reset for core cols. 15:0	0
			Funtions: (T)rigger and timing, (I)nput/ouput, (C)alibration, (M)asking, () Other		
T	47	TriggerConfig	1,9	Trigger mode, Latency (9)	0,500
T	48	SelfTriggerConfig_1	1,1,4	Self Trig.: En., ToT thresh.: En., value (9.4)	0,1,1
T	49	SelfTriggerConfig_0	10,5	Self Trig.: delay, multiplier (9.4)	100,1
T	50	SelfTriggerDeadTime	16	inhibit after each trig. (BXs 0=off)	0
T	51	HitOrPatternLUT	16	Self Trig. HitOR logic program (9.4)	0
T	* 52	ReadTriggerConfig	2,12	Col. Read delay, Read Trig. Decision time (BXs)	0,1000
T	53	TruncationTimeoutConf	12	Event truncation timeout (BXs, 0=off)	0
C	* 54	CalibrationConfig	1,1,6	Cal injection: Ana/Dig (0/1), Ana mode, fine delay	0,1,0
T	55	CLK_DATA_FINE_DELAY	6,6	Fine delays for Clock, Data (14)	0,0
C	56	VCAL_HIGH	12	VCAL high level	500
C	57	VCAL_MED	12	VCAL medium level	300
C	* 58	MEAS_CAP	1,1,1	Cap Meas: En. Par, En.; VCAL range bit	0,0,0
I	59	CdrConf	1,3	CDR: phase det sel., CLK sel. (14)	0,0,0
I	60	ClkTriplConf	3,3,3,3	Trpl. En: 40MHz, 160MHz, Data Merger, Aurora	8,8,8,8
I	61	ChSyncConf	5	Chan. Synch. Lock Thresh. (unlock is ×2)	16
	* 62	GlobalPulseConf	16×1	Global pulse routing	0
	63	GlobalPulseWidth	9	Global Pulse Width (in BX, 0=1)	1
I	64	ServiceDataConf	1,8	Service block En., Periodicity N_D (10.2)	0,50
	65	ToTConfig	1,1,1,1,9	En: PToT, PToA, 80MHz, 6b to 4b; PToT Latency	0,0,0,0,0
M	66	PrecisionToTEnable_3	6×1	Enable PToT for core cols. 53:48	0
M	67	PrecisionToTEnable_2	16×1	Enable PToT for core cols. 47:32	0
M	68	PrecisionToTEnable_1	16×1	Enable PToT for core cols. 31:16	0
M	69	PrecisionToTEnable_0	16×1	Enable PToT for core cols. 15:0	0
I	* 70	DataMerging	4,1,1,1,4,1,1	Invert(4), Ch.ID, 1.28CK gate, CK sel, En.(4), Ch.bond, GPO sel.	0,1,1,0,0,0,1
I	71	DataMergingMux	8×2	Input and Output Lane mapping	3,2,1,0,3,2,1,0
M	72	EnCoreColumnCalibration_3	6×1	CAL enable for core cols. 53:48	6×1
M	73	EnCoreColumnCalibration_2	16×1	CAL enable for core cols. 47:32	16×1
M	74	EnCoreColumnCalibration_1	16×1	CAL enable for core cols. 31:16	16×1
M	75	EnCoreColumnCalibration_0	16×1	CAL enable for core cols. 15:0	16×1
I	76	DataConcentratorConf	1,1,1,8	n/u: CRC, BCID, L1ID; evts/stream-1	0,0,0,0
I	* 77	CoreColEncoderConf	1,1,9,3,1,1	Drop ToT, raw map, MaxHits, MaxToT, A, B	0,0,0,0,0,1
I	78	EnMaxHitsLimit	6×1	Drop hits above MaxHits for cols. 53:48	6×0
I	79	EnMaxHitsLimit	16×1	Drop hits above MaxHits for cols. 47:32	16×0

Table 22: Global configuration register main table. (*) indicates further details are given following the table. Section references are given in (). The Size column may list multiple fields by their size in bits. The rightmost value is always the field starting at register bit [0] (i.e. all words are little-endian). Thus, the numbers "1,3,2" in the Size column would indicate a word with 3 fields: [5],[4:2],[1:0]. n/u Stands for not used in ATLAS chip.

	Addr.	Name	Size (bits)	Description	CMS Default
I	80	EnMaxHitsLimit	16×1	Drop hits above MaxHits for cols. 31:16	16×0
I	81	EnMaxHitsLimit	16×1	Drop hits above MaxHits for cols. 15:0	16×0
I	82	EnIHR	6×1	Drop isolated hits below MaxTOT for cols. 53:48	6×0
I	83	EnIHR	16×1	Drop isolated hits below MaxTOT for cols. 47:32	16×0
I	84	EnIHR	16×1	Drop isolated hits below MaxTOT for cols. 31:16	16×0
I	85	EnIHR	16×1	Drop isolated hits below MaxTOT for cols. 15:0	16×0
I	86	EvenMask	16	Isolated hit filter mask: Even cols.	0
I	87	OddMask	16	Isolated hit filter mask: Odd cols.	0
	88	EfusesConfig	16	Efuses En. (to Read set 0F0F, to Write set F0F0)	0
	89	EfusesWriteData1	16	Data to be written to Efuses (1 of 2)	0
	90	EfusesWriteData0	16	Data to be written to Efuses (2 of 2)	0
I	91	PhaseDetectorConfig	1,4,8	DataMerg phase manual override, mode(4), choice(8)	0,0,0
I	92	AuroraConfig	1,4,6,2	AURORA: Alt. Out, En. PRBS, En. Lanes, CCWait, CCSend	0,0,15,25,3
I	93	AURORA_CB_CONFIG1	8	Aurora Chann. Bonding Wait [19:12]	255
I	94	AURORA_CB_CONFIG0	12,4	Aurora Chann. Bond Wait [11:0], CBSend	4095,0
I	95	AURORA_INIT_WAIT	11	Aurora Initialization Delay	32
I	96	AURORA_ALT_OUT1	4	Aurora alt. output reg. 1	0
I	97	AURORA_ALT_OUT0	16	Aurora alt. output reg. 0	0
I	98	OUTPUT_PAD_CONFIG	4,1,1,4,3	GP_CMOS: pattern, En, DS, GP_LVDS: Enables, strength	5,1,0,15,7
I	99	GP_CMOS_ROUTE	6	GP_CMOS MUX select (16.4)	13
I	100	GP_LVDS_ROUTE_1	6,6	GP_LVDS(3), GP_LVDS(2) MUX select (16.4)	14,10
I	101	GP_LVDS_ROUTE_0	6,6	GP_LVDS(1), GP_LVDS(0) MUX select (16.4)	2,0
I	102	DAC_CP_CDR	10	CDR CP Bias (values <15 are set to 15)	40
I	103	DAC_CP_FD_CDR	10	CDR FD CP bias (values <100 are set to 100)	400
I	104	DAC_CP_BUFF_CDR	10	CDR unity gain buffer bias	200
I	105	DAC_VCO_CDR	10	CDR VCO bias (values <700 are set to 700)	1023
I	106	DAC_VCOBUFF_CDR	10	CDR VCO buffer bias (values <200 are set to 200)	500
I	107	SER_SEL_OUT	4×2	CML 3-0 content. 0=CK/2, 1=AURORA, 2=PRBS7, 3=0	1,1,1,1
I	108	CML_CONFIG	2,2,4	CML out: Inv. Tap 2,1; En. Tap 2,1; En. Lane 3,2,1,0	0,0,4x1
I	109	DAC_CML_BIAS_2	10	CML drivers tap 2 amplitude (pre-emph)	0
I	110	DAC_CML_BIAS_1	10	CML drivers tap 1 amplitude (pre-emph)	200
I	111	DAC_CML_BIAS_0	10	CML drivers tap 0 amplitude (main)	900
				Monitoring and Test	
	112	MonitorConfig	1,6,6	Monitor pin: En., I. MUX sel., V. MUX sel.	0,63,63
	113	ErrWngMask	8×1	Error and Warning Message disable Mask	0
	114	MON_SENS_SLDO	1,4,1,1,4,1	Tsense LDO: En.A, DEM, Bias, En.D, DEM, Bias (12.5)	0,0,0,0,0,0
	115	MON_SENS_ACB	1,4,1	Tsense center: En., DEM, Bias (12.5)	0,0,0
	116	MON_ADC	1,1,1,6	Vref for Rsense: bot., top.; Vref in; ADC trim bits	0,0,1,0
	117	DAC_NTC	10	Current output DAC for the external NTC	100
M	118	HITOR_MASK_3	6×1	HitOR disable for core cols. 53:48	0
M	119	HITOR_MASK_2	16×1	HitOR disable for core cols. 47:32	0
M	120	HITOR_MASK_1	16×1	HitOR disable for core cols. 31:16	0
M	121	HITOR_MASK_0	16×1	HitOR disable for core cols. 15:0	0

Table 22: Global configuration register main table. (*) indicates further details are given following the table. Section references are given in (). The Size column may list multiple fields by their size in bits. The rightmost value is always the field starting at register bit [0] (i.e. all words are little-endian). Thus, the numbers "1,3,2" in the Size column would indicate a word with 3 fields: [5],[4:2],[1:0]. n/u Stands for not used in ATLAS chip.

	Addr.	Name	Size (bits)	Description	CMS Default
I	122	AutoRead0	9	Auto-Read register address A for lane 0	134
I	123	AutoRead1	9	Auto-Read register address B for lane 0	135
I	124	AutoRead2	9	Auto-Read register address A for lane 1	137
I	125	AutoRead3	9	Auto-Read register address B for lane 1	138
I	126	AutoRead4	9	Auto-Read register address A for lane 2	140
I	127	AutoRead5	9	Auto-Read register address B for lane 2	151
I	128	AutoRead6	9	Auto-Read register address A for lane 3	152
I	129	AutoRead7	9	Auto-Read register address B for lane 3	156
	* 130	RingOscConfig	15×1	Ring oscillator enable bits (13.6)	1,5×0,1,8×0
	131	RingOscRoute	3,6	Select which RO to read from block A, B (13.6)	0,0
	132	RING_OSC_A_OUT	16	Ring oscillator block A output (rd. only) (13.6)	n/a
	133	RING_OSC_B_OUT	16	Ring oscillator block B output (rd. only) (13.6)	n/a
	134	BCIDCnt	16	Bunch counter (rd. only)	n/a
	135	TrigCnt	16	Received trigger counter (rd. only)	n/a
	136	ReadTrigCnt	16	Received or internal ReadTrigger ctr (rd. only)	n/a
	137	LockLossCnt	16	Channel Sync lost lock counter (rd. only)	n/a
	138	BitFlipWngCnt	16	Bit Flip Warning counter (rd. only)	n/a
	139	BitFlipErrCnt	16	Bit Flip Error counter (rd. only)	n/a
	140	CmdErrCnt	16	Command Decoder error message ctr (rd. only)	n/a
	141	RdWrFifoErrorCount	16	Writes and Reads when fifo was full ctr (rd. only)	n/a
	142	AI_REGION_ROW	9	Auto Increment current row value (rd. only)	n/a
	143	HitOr_3_Cnt	16	HitOr_3 Counter (rd. only)	n/a
	144	HitOr_2_Cnt	16	HitOr_2 Counter (rd. only)	n/a
	145	HitOr_1_Cnt	16	HitOr_1 Counter (rd. only)	n/a
	146	HitOr_0_Cnt	16	HitOr_0 Counter (rd. only)	n/a
	147	GatedHitOr_Cnt_1	8,8	Counters for gated HitOr's (rd. only)	n/a
	148	GatedHitOr_Cnt_0	8,8	Counters for gated HitOr's (rd. only)	n/a
	149	Pixel_SEU_Cnt	16	Counts pixel reg. bit flips (rd. only)	n/a
	150	GlobalConfig_SEUCnt	4,12	Counters for global config single bit flips (rd. only)	n/a
	151	SkippedTriggerCnt	16	Skipped Trigger counter (rd. only)	n/a
	* 152	DataDecodingVals	16	Encodes selected data output format options (rd. only)	n/a
	153	EfusesReadData1	16	Readback of efuses 1 of 2 (rd only)	n/a
	154	EfusesReadData0	16	Readback of efuses 2 of 2 (rd only)	n/a
	* 155	MonitoringDataADC	12	ADC value (rd. only)	n/a
	156	PadReadout	4,4	Wire bond pad values: Chip ID, Iref Trim (rd only)	n/a
	157-159	SEU00-SEU02	16	GR dummies for SEU meas. Full protection	0
	160-191	SEU_nodelxx (xx=0-31)	16	GR dummies for SEU meas without 3-phase clocks	0
	192-255	SEU_noTMR (xx=0-63)	16	Dummies for SEU meas without triple redundancy	0

*** 0 PIX_PORTAL:** The pixel portal allows access the the registers within any pixel. Every pixel has 8 configuration bits:

Bits	Name	Description
[0]	Enable	Include the pixel in the DAQ data path
[1]	Cal Enable	Turn on charge injection (*)
[2]	HitOr Enable	Add the pixel to its wired OR core col. hit line
[3:6]	TDAC value	Value for in-pixel theshold trim DAC
[7]	TDAC sign	Selects differential branch set to TDAC value

Table 23: ATLAS pixel configurations bits. (*) Whether charge injection is digital or analog is controlled by global configuration register CalibrationConfig. The 5-bit pixel TDAC is made up of a 4-bit value and a sign bit.

*** 3 PIX_MODE:** Consisis of: EnSEUCount to enable counting of SEU events in pixel registers, HitSampleMode bit to sample pixel outputs asnchronlusly (0) or synchronously (1), Broadcast enable bit, the ConfWrConfig bit, and the AutoRow enable bit. The Broadcat enable bit causes the column pair value of Reg. 1 to be ignored and all pixel in the same row to be written with the given value. The ConfWrConfig bit is only active when using the Write Multiple mode of the write register command. In this mode only 10 bits are available in for two pixels in each write cycle, and they are assigned according to this bit. ConfWrConfig = 1 writes the Pixel Config (5-bits per pixel) and ConfWrConfig = 0 writes FrontEnd Config (3-bits per pixel). The bits that are not written will remain unchanged. Auto Row mode is described in Sec. 8.8.

*** 4,5 PIX_DEFAULT:** A special pair of registers that control the multiplexers for pixel configuration. When the correct value is programmed in both registers the programmed configuration (which must first be written) will be used. Until then the hard-wired default configuration will be active. Any single bit flip in the correct value of each register will still be interpreted as the correct value.

*** 6,7 GCR_DEFAULT:** A special pair of registers that control the multiplexers for global configuration. When the correct value is programmed in both registers the programmed configuration will be used. Until then the hard-wired default configuration will be active. Any single bit flip in the correct value of each register will still be interpreted as the correct value. In this case the default configuration will also be present in the registers soon after the chip clock is present, so the user need not program anything before switching (See 3).

*** 37 LEAKAGE_FEEDBACK:** Despite its address after Lin FE biases, it enables the Diff. FE leakage current compensation (bit 0) and a second feedback capacitance to reduce the gain (bit 1).

*** 43-46 EnCoreColumnReset:** In the pixel matrix the clocks are gated unless there is hit activity. However, upon power up, the gating state is arbitrary and enabling the clock to the the columns can result in significant power consumption (up to double normal power). Therefore, a reset is provided to that columns can be reset to their full clock gated state as they have their clock enabled. The EnCoreColumnReset registers gate the reset signals to the columns, so that only a limited number of columns can be reset at the same time, to avoid a large current spike. It is recommended to set enable EnCoreCol and EnCoreColumnReset equal to each other, and at start of operation cycle though selecting a small group of core columns at a time and issuing a reset signal for each

group. The reset signal is issued by the Clear command (Sec. 8.2.1). After this initialization all core columns can be enabled.

*** 52 ReadTriggerConfig:** The column read delay is the number of BX clocks allowed for the read token to propagate and data valid to appear at the bottom of column. The default (0) allows two clocks. This should normally not be changed but gives the option to allow more time in case of slower than expected propagation, especially after irradiation. The read trigger decision time is only relevant in 2-trigger mode. It is effectively the L0 to L1 latency. It is the number of bunch clocks allowed after a L0 trigger before readout or clear action is taken. A read trigger command must arrive before this delay for a readout action to be taken (See 9).

*** 54 CalibrationConfig:** Injection can be either analog or digital, selected by bit [7]. Bit [6] selects the mode of analog injection, which can be regular (same for every pixel, default) or alternating (see Sec. 6)

*** 58 MEAS_CAP:** Contains the enable bits for the capmeasure circuit (see Sec. 13.8), as well as the VCAL range bit (see Sec. 6). If VCAL range is zero the injection voltage full scale is from 0 to Vref_ADC/2, if 1 it is from 0V to Vref_ADC. Note Vref_ADC is used as a reference for the calibration injection voltage as well as the Generic ADC.

*** 62 GlobalPulseConf:** The global pulse allows to toggle internal signals within the chip.

Bit	Route to:	Bit	Route to:
0	Reset Channel Synchronizer	8	Reset Data Serializers
1	Reset Command Decoder	9	Reset ADC
2	Reset Global Configuration	10	Reset e-fuses
3	Reset Aurora only	11	Send Cal Reset pulse
4	Reset Data Path (not Aurora)	12	ADC StartOfConversion
5	CLEAR also resets Aurora	13	Send Start Signal to Ring Oscillators block A
6	Reset BCID, LV1ID and ReadTrigger cntrs	14	Send Start Signal to Ring Oscillators block B
7	CLEAR also resets above cntrs	15	Send Start Signal to EfusesProgrammer

Table 24: Global pulse routing choices. Any number of bits can be selected simultaneously.

*** 70 DataMerging:** Switch actions are: Invert the polarity of input signals to DataMerger , Enable sending Chip ID in output data, Enable Gating of 1280 MHz Data Merge clock (0 enables deserializer, 1 disables it), Select which clock to use for data merging (0=640 Mz 1=1280 MHz), Select which of the Data Mergers are enabled, selects if first two inputs are channel bonded (to merge one secondary chip at 640 Mbps), The final bit (GPOsel) is a detalied debugging feature for testing only. Set to 1 it sends to the General Purpose Output (GPO) the deserialised data at the output of the phase detector for channel 0. Set to 0 it sends instead the output coming from the deserialiser in the pll block (also for channel 0). As this block outputs 16 bits of data sampled 4 times, bits 12,8,4 and 0 will appear on the GPO, which correspoind to one of the phases of the phase detector.

*** 77 Data Encoder Config:** The last two bits, labeled A and B, are place holders for isolated hit removal functionality that was not implemented. They are included becasue they offset the position of the remaining bits. Raw map turns off compression, so that every hit map is exactly 16 bits. Max Hits sets the column readout truncation threshold, where the value 0 deisables truncation.

*** 130 RingOscConfig:** Enable bits for: Ring Osc. B Clear, Ring Osc. B Enable BL, Ring Osc. B Enable BR, Ring Osc. B Enable CAPA, Ring Osc. B Enable FF, Ring Osc. B Enable LVT, Ring Osc. A Clear, Ring Osc. A Enable[7:0].

*** 152 DataDecodingVals:** This register provides summary information to decode the output data.

Bit	Meaning	Reg.	Bit	Meaning	Reg.	Bit	Meaning	Reg.
15	Sample mode (n/u)	3	9	Include LV1 ID (n/u)	76	4	En Service Data	64
14	ToT 80MHz Count	65	8	Drop TOT (binary read)	77	3	En Prec. ToA	65
13	ToT 6 to 4 Mapping	65	7	Raw bit map	77	2	En Prec. ToT	65
12	Include Chip ID	70	6	Events per Stream -1	76	1	En col hits limit	78-81
11	Add CRC bits (n/u)	76	5	Self Trigger	48	0	En Iso Hit removal	82-85
10	Include BCID (n/u)	76						

Table 25: Bits of data ouptut format decoding register. Reading this register the DAQ/offline can always find out exactly what endoding options were used. This information is needed in order to decode the data. It can also be obtained from the chip configuration used, but this puts it all in one place. Features not used in ATLAS chip are marked n/u, but the bits still will be set or not according to the configuration registers for those features.

*** 155 MonitoringDataADC:** This value is updated every time the ADC performs a conversion, which is triggered by a global pulse. Reading the register multiple times between ADC conversions will return the same value. Reading the register does not trigger a conversion.

16.3 IMUX and VMUX selection values

Setting	Selected Input	Setting	Selected Input	Setting	Selected Input
0	IREF main ref. current	11	Capmeasure parasitic	22	DIFF FE Preamp Top-Left
1	CDR VCO main bias	12	DIFF FE Preamp Main array	23	DIFF FE VTH1 Right
2	CDR VCO buffer bias	13	DIFF FE PreComp	24	DIFF FE Preamp Top
3	CDR CP current	14	DIFF FE Comparator	25	DIFF FE Preamp Top-Right
4	CDR FD current	15	DIFF FE VTH2	26	not used
5	CDR buffer bias	16	DIFF FE VTH1 Main array	27	not used
6	CML driver tap 2 bias	17	DIFF FE LCC	28	Ana. input current/21000
7	CML driver tap 1 bias	18	DIFF FE Feedback	29	Ana. shunt current/21600
8	CML driver main bias	19	DIFF FE Preamp Left	30	Dig. input current/21000
9	NTC_pad current	20	DIFF FE VTH1 Left	31	Dig. shunt current/21600
10	Capmeasure circuit	21	DIFF FE Preamp Right	32-62	not used
				63	high Z

Table 26: Current multiplexer (I_mux) assignments for ATLAS chip.

Setting	Selected Input	Setting	Selected Input	Setting	Selected Input
0	Vref_ADC (GADC)	10	DIFF FE VTH1 Main array	31	Vref_CORE
1	I_mux pad voltage	11	DIFF FE VTH1 Left	32	Vref_PRE
2	NTC_pad voltage	12	DIFF FE VTH1 Right	33	VINA / 4
3	VCAL_DAC / 2 (Sec. 6.3)	13	RADSENS Ana. SLDO	34	VDDA / 2
4	VDDA / 2 from capmeasure	14	TEMPSENS Ana. SLDO	35	VrefA
5	Poly TEMPSENS top	15	RADSENS Dig. SLDO	36	VOFS / 4
6	Poly TEMPSENS bottom	16	TEMPSENS Dig. SLDO	37	VIND / 4
7	VCAL_HI	17	RADSENS center	38	VDDD / 2
8	VCAL_MED	18	TEMPSENS center	39	VrefD
9	DIFF FE VTH2	19-30	Ana. GND	40-62	not used
				63	high Z

Table 27: Voltage multiplexer (V_mux) assignments for ATLAS chip.

16.4 General Purpose LVDS and CMOS Output Assignments

The same selection codes apply to all outputs. Each output has a dedicated multiplexer and there is no issue selecting the same signal for multiple outputs.

#	Selected Signal	Def	#	Selected Signal
0	Command input to chip (after LVDS receiver)	LVDS0	32	Logic combination of HitOr's from SelfTrigger
1	Inverted Command input ot chip		33	CalEdge output from Command Decoder
2	Recovered command data before applyng delay	LVDS1	34	CalAux output from Command Decoder (CMD)
3	Inverted Recovered command data before delay		35	GlobalPulse output from Command Decoder
4	Recovered command data after applyng delay		36	skipped trigger Warning from CMD
5	Inverted Recovered command data after delay		37	Error signal from Command Decoder
6	recovered 160 MHz clock before applying delay		38	Error counter increment from CMD
7	recovered 160 MHz clock after applying delay		39	SEU Error signal from Command Decoder
8	40MHz Clock to Pixel Matrix		40	SEU Error counter increment from CMD
9	640MHz Clock to Precision ToT		41	SEU Warning from Command Decoder
10	Power on Reset signal	LVDS2	42	SEU Warning counter increment from CMD
11	CMD Idle. Goes low if no transitions detected on CMD input, which causes a reset (*)		43	Channel Synchronizer Lock Lost signal
12	Same as 11 but synchronized to clock		44	GlobalConfiguration SEU_full counter incr.
13	Channel Synchronizer Lock signal	CMOS	45	GlobalConfiguration SEU_single counter incr.
14	PLL Lock signal	LVDS3	46	Bit 3 from pattern in OUTPUT_PAD_CONFIG (reg. 98)
15	Spies on CDR/PLL up signal		47	Bit 2 from pattern in OUTPUT_PAD_CONFIG (reg. 98)
16	Spies on CDR/PLL down signal		48	Bit 1 from pattern in OUTPUT_PAD_CONFIG (reg. 98)
17	Spies on CDR/PLL up_fd signal		49	Bit 0 from pattern in OUTPUT_PAD_CONFIG (reg. 98)
18	Spies on CDR/PLL down_fd signal		50	End of ADC conversion flag
19	Data arriving at Data Merger input 3		51	LaneUp signal from Aurora Ch. 0
20	Data arriving at Data Merger input 2		52	OR of all Lane Up signals before the switch matrix
21	Data arriving at Data Merger input 1		53	OR of all DeserializedValid signals of LaneUnit
22	Data arriving at Data Merger input 0		54	OR of all SyncDataValid signals of LaneUnit
23	Low power mode indicator (low = low pwr)		55	ChannelUp signal from DeserializerAuroraBonder
24	Trigger signal from Command Decoder		56	Fifo empty sig. in DesData block of ch.0
25	ReadTrigger signal from Command Decoder		57	Fifo empty in DesMonitor of ch.0
26	Trigger pulse from Self Trigger block to CMD		58	Bit 3 of GpoSelected word
27	Read Trigger pulse from Self Trigger to CMD		59	Bit 2 of GpoSelected word
28	HitOr number 3		60	Bit 1 of GpoSelected word
29	HitOr number 2		61	Bit 0 of GpoSelected word
30	HitOr number 1		62	FoundTrans signal from LaneUnit of ch.0
31	HitOr number 0		63	"ChosenâĂŹâĂŹ signal from LaneUnit of ch.0

Table 28: Signal source selection for GP_LVDS and GP_CMOS outputs. Each of the 4 LVDS and single CMOS outputs is indpendently assigned a source. The default column shows which of the outputs has that source as its default. (*) The synchronized CMD idle signal (12) resets the chip. Thus, idling the command input can be used as a hard reset with no need to power cycle.

16.5 Internal and External Component Nominal Values

Component	Type	Value	Tolerance	Reference
Injection capacitor	MOM	8.02 fF	±10%	Fig. 32
Diff FE feedback cap (high gain)	parasitic	3.73 fF	n/a	Fig. **??**
Diff FE feedback cap (low gain)	parasitic	3.28 fF	n/a	Fig. **??**
Resistive temp. sensor	p^+ Poly w/silicide	10 K	±30%	Sec. 12.4
Resistive temp. reference	p^+ Poly w/o silicide	16 K	±12.5%	Sec. 12.4
Internal receiver terminations	Poly	100 Ω	±12.5%	Sec. 8.1
Chip ID pull up resistors	Poly	40 k	±12.5%	to VDDD
IREF trim pull up resistors	Poly	40 k	±12.5%	to VDD_PRE

Table 29: Internal passive component types and values.

Component	Function	Value	Reference	Notes
R_{Iref}	Current reference resistor	22.6 K	Fig. 15	V_{BGR} = 450 mV makes Iref = 20 μA
R_{VrefA}	Sets analog regulator ref. V.	30 K	Fig. 15	VrefA = 0.6 V baseline
R_{VrefD}	Sets analog regulator ref. V.	30 K	Fig. 15	VrefD = 0.6 V baseline
R_{shuntA}	Sets shunt slope	400 Ω	Figs. 13, 14	V_{IN}=1.4 V at 1 A
R_{shuntD}	Sets shunt slope	400 Ω	Figs. 13, 14	V_{IN}=1.4 V at 1 A
R_{OFS1}	Sets the shunt offset V.	24.9 K	Figs. 13, 15	VOFS/2 = 0.5 V baseline
R_{OFS2}	Sets the low power offset V.	10 K	Fig. 15	VOFS/2 = 0.7 V for low power
R_{IMUX}	Converts monitoring I output to a voltage	4.99 K		
$R_{VrefADC}$	Sets the ADC reference voltage	84.5 K		VREF_ADC = 845 mV baseline
R_{CMD}	termination resistor	100 Ω		
C_{VDDA}	Regulator output bypass	≥0.55 μF	Fig. 8	4 instances (2.2 μF total)
C_{VDDD}	Regulator output bypass	≥0.55 μF	Fig. 8	4 instances (2.2 μF total)
C_{VPRE}	Pre-regulator output bypass	≥0.45 μF		
C_{VDD_PLL}	Phase locked loop bypass	2.2 μF		
C_{VDD_CML}	Output driver bypass	2.2 μF		2 instances
C_{VINA}	Regulator input cap	≥6.8 μF	Fig. 8	TBD by SP chain operation
C_{VIND}	Regulator input cap	≥6.8 μF	Fig. 8	TBD by SP chain operation
C_{VrefA}	Vref bypass	100 nF		
C_{VrefD}	Vref bypass	100 nF		
C_{IREF}	Iref bypass	100 nF		
$C_{VrefADC}$	Vref bypass	100 nF		
C_{IMUX}	filter for current mux	22 nF		only for wafer probing
C_{IVMUX}	filter for voltage mux	22 nF		optional
C_{CMD}	command A/C coupling	10-100 nF		2 instances
$C_{LP\_EN\_AC}$	Low power mode enable A/C coupling	100 nF	Fig. 19	1 instance

Table 30: External passive component types and values.

16.6 Command and Trigger Encoding

Command	Encoding		(T)ag, (A)ddress or (D)ata 5-bit content					
Sync	1000_0001	0111_1110						
PLLlock	1010_1010	1010_1010						
Trigger	tttt_tttt	Tag[0..53]						
Clear	0101_1010	ID<4:0>						
Global Pulse	0101_1100	ID<4:0>						
Cal	0110_0011	ID<4:0>	D<19:15>	D<14:10>	D<9:5>	D<4:0>		
WrReg(0)	0110_0110	ID<4:0>	0,A<8:5>	A<4:0>	D<15:11>	D<10:6>	D<5:1>	D<0>,0000
WrReg(1)	0110_0110	ID<4:0>	1,xxxx	xxxxx	N×(D<9:5>	D<4:0>)		
RdReg	0110_0101	ID<4:0>	0,A<8:5>	A<4:0>				
Read_trigger(*)	0110_1001	ID<4:0>	00,T<7:5>	T<4:0>				

Table 31: This is a duplicate of Table 6. List of protocol commands/frames and address or data fields associated with each. Unused padding bits are indicated by "0". Double vertical lines denote frame boundaries. tttt_tttt is one of 15 trigger commands (Table 7). The before-encoded bit content of chip ID, Address or Data is shown. These are all encoded as 8-bit data symbols (Table 32). (*) Read_trigger is a legacy command and should not be used in RD53C, as the trigger mode requiring it has been deprecated.

Symbol Name	Encoding	Data Value	Symbol Name	Encoding	Data Value
Data_00	0110_1010	5'b00000	Data_16	1010_0110	5'b10000
Data_01	0110_1100	5'b00001	Data_17	1010_1001	5'b10001
Data_02	0111_0001	5'b00010	*Data_18*	*0101_1001*	5'b10010
Data_03	0111_0010	5'b00011	Data_19	1010_1100	5'b10011
Data_04	0111_0100	5'b00100	Data_20	1011_0001	5'b10100
Data_05	1000_1011	5'b00101	Data_21	1011_0010	5'b10101
Data_06	1000_1101	5'b00110	Data_22	1011_0100	5'b10110
Data_07	1000_1110	5'b00111	Data_23	1100_0011	5'b10111
Data_08	1001_0011	5'b01000	Data_24	1100_0101	5'b11000
Data_09	1001_0101	5'b01001	Data_25	1100_0110	5'b11001
Data_10	1001_0110	5'b01010	Data_26	1100_1001	5'b11010
Data_11	1001_1001	5'b01011	Data_27	1100_1010	5'b11011
Data_12	1001_1010	5'b01100	Data_28	1100_1100	5'b11100
Data_13	1001_1100	5'b01101	Data_29	1101_0001	5'b11101
Data_14	1010_0011	5'b01110	Data_30	1101_0010	5'b11110
Data_15	1010_0101	5'b01111	Data_31	1101_0100	5'b11111

Table 32: List of command symbols used to encode data values. All symbols are the same as in RD53A except for Data_18, which is shown in italics. The RD53A Data_18 symbol is now the PLLlock command.

Symbol Name	Encoding	Trigger Pattern	Symbol Name	Encoding	Trigger Pattern
			Trigger_08	0011_1010	T000
Trigger_01	0010_1011	000T	Trigger_09	0011_1100	T00T
Trigger_02	0010_1101	00T0	Trigger_10	0100_1011	T0T0
Trigger_03	0010_1110	00TT	Trigger_11	0100_1101	T0TT
Trigger_04	0011_0011	0T00	Trigger_12	0100_1110	TT00
Trigger_05	0011_0101	0T0T	Trigger_13	0101_0011	TT0T
Trigger_06	0011_0110	0TT0	Trigger_14	0101_0101	TTT0
Trigger_07	0011_1001	0TTT	Trigger_15	0101_0110	TTTT

Table 33: This is a duplicate of Table 7. List of trigger symbols used to encode the 15 possible trigger patterns spanning four bunch crossings. Note there is no 0000 pattern as that is the absence of an trigger. The Trigger_01 (000T) means that the first bunch crossing of the trigger window is meant to be readout, and the extended tag returned will have 00 following the supplied tag base.

Tag base	Extended tag range	Symbol code	Symbol name	Tag base	Extended tag range	Symbol code	Symbol name
0	0-3	0110_1010	Data_00	27	108-111	1100_1010	Data_27
1	4-7	0110_1100	Data_01	28	112-115	1100_1100	Data_28
2	8-11	0111_0001	Data_02	29	116-119	1101_0001	Data_29
3	12-15	0111_0010	Data_03	30	120-123	1101_0010	Data_30
4	16-19	0111_0100	Data_04	31	124-127	1101_0100	Data_31
5	20-23	1000_1011	Data_05	32	128-131	0110_0011	Cal
6	24-27	1000_1101	Data_06	33	132-135	0101_1010	Clear
7	28-31	1000_1110	Data_07	34	136-139	0101_1100	GlobalPulse
8	32-35	1001_0011	Data_08	35	140-143	1010_1010	PllLock
9	36-39	1001_0101	Data_09	36	144-147	0110_0101	ReadReg
10	40-43	1001_0110	Data_10	37	148-151	0110_1001	ReadTrigger
11	44-47	1001_1001	Data_11	38	152-155	0010_1011	Trigger_01
12	48-51	1001_1010	Data_12	39	156-159	0010_1101	Trigger_02
13	52-55	1001_1100	Data_13	40	160-163	0010_1110	Trigger_03
14	56-59	1010_0011	Data_14	41	164-167	0011_0011	Trigger_04
15	60-63	1010_0101	Data_15	42	168-171	0011_0101	Trigger_05
16	64-67	1010_0110	Data_16	43	172-175	0011_0110	Trigger_06
17	68-71	1010_1001	Data_17	44	176-179	0011_1001	Trigger_07
18	72-75	0101_1001	Data_18	45	180-183	0011_1010	Trigger_08
19	76-79	1010_1100	Data_19	46	184-187	0011_1100	Trigger_09
20	80-83	1011_0001	Data_20	47	188-191	0100_1011	Trigger_10
21	84-87	1011_0010	Data_21	48	192-195	0100_1101	Trigger_11
22	88-91	1011_0100	Data_22	49	196-199	0100_1110	Trigger_12
23	92-95	1100_0011	Data_23	50	200-203	0101_0011	Trigger_13
24	96-99	1100_0101	Data_24	51	204-207	0101_0101	Trigger_14
25	100-103	1100_0110	Data_25	52	208-211	0101_0110	Trigger_15
26	104-107	1100_1001	Data_26	53	212-215	0110_0110	WrReg

Table 34: Tag base codes. All 8-bit symbols are re-used to provide the maximum number of tag bases.

16.7 Output Tags

Tag values (decimal)	**Meaning**
0-215	extended tags from trigger command
216-219	Single bit-flip detected in tag symbol of a trig. command
220-223	Unrecognized tag symbol
224-255	Self-trigger tag values

Table 35: This is a copy of Table 11. Possible extended tag values and their meaning.

16.8 ToT Table

Output	**True ToT bin (low edge) [BX]**			
4-bit	**40 MHz speed**		**80 MHz speed**	
code	**4-bit (DEF)**	**6-to-4 bit**	**4-bit**	**6-to-4 bit**
0	0	0	0	0
1	1	1	0.5	0.5
2	2	2	1	1
3	3	3	1.5	1.5
4	4	4	2	2
5	5	5	2.5	2.5
6	6	6	3	3
7	7	7	3.5	3.5
8	8	8	4	4
9	9	12	4.5	6
10	10	16	5	8
11	11	20	5.5	10
12	12	24	6	12
13	13	28	6.5	14
14	$\geq$14	$\geq$32	$\geq$7	$\geq$16

Table 36: This is a copy of Table 5. True ToT value in bunch crossing (BX = 25 ns units) for each output ToT 4-bit code, depending on speed (40 or 80 MHz) and compression (4 bit or 6-to-4 bit) settings. Always the low edge of the true ToT bin is shown. For example code 3 having a true ToT low edge of 3 means the true ToT was at least 3 bunch crossings and at most x, where x is the true ToT low edge of the next code (4 in this case). The last bin (code 14) has no high edge and includes all overflows. Code 15 means "no hit" and should never be seen because unhit pixels are internally suppressed.

16.9 Ring Oscillator Assignments

ROSC Nbr.	Type	Len.	ROSC Nbr.	Type	Len.
0	Strgth. 0 inv. clk. drvr.	55	4	Strgth. 0 4-input NAND	19
1	Strgth. 4 inv. clk. drvr.	51	5	Strgth. 4 4-input NAND	19
2	Strgth. 0 inverter	55	6	Strgth. 0 4-input NOR	19
3	Strgth. 4 inverter	51	7	Strgth. 4 4-input NOR	19

Table 37: Bank A ring oscillator types and lengths (in number of gates). The given lengths result in a typical frequency of about 600 MHz before irradiation. Each oscillator has its own Enable bit.

ROSC Nbrs.	Type	Eff. Len.	Group
0 & 1	Strgth. 0 inv. clk. driver	38.2	B-left, B-right
2 & 3	Strgth. 4 inv. clk. driver	44.5	B-left, B-right
4 & 5	Strgth. 0 inverter	38.1	B-left, B-right
6 & 7	Strgth. 4 inverter	44.3	B-left, B-right
8 & 9	Strgth. 0 4-input NAND	12.6	B-left, B-right
10 & 11	Strgth. 4 4-input NAND	16	B-left, B-right
12 & 13	Strgth. 0 4-input NOR	14.5	B-left, B-right
14 & 15	Strgth. 4 4-input NOR	14.5	B-left, B-right
16 & 17	Strgth. 0 scan D-flip-flop	6.1	FF
18 & 19	Strgth. 1 D-flip-flop	6.2	FF
20 & 21	Strgth. 1 Neg. edge D-flip-flop	5	FF
22	Strgth. 0 LVT inverter	40.6	LVT
23	Strgth. 4 LVT inverter	56	LVT
24	Strgth. 0 LVT 4-input NAND	16.5	LVT
25	Strgth. 4 LVT 4-input NAND	22.8	LVT
26-33	Strgth. 4 inj-cap-loaded 4-input NAND	16.8	CAPA

Table 38: Bank B ring oscillator types and lengths (in number of equivalent gates). The lengths given result in a typical frequency of about 800 MHz before irradiation. The oscillators are connected in separately enabled groups (there are no individual enable bits for each oscillator)

A. Aurora 64b66b Technical Reference

References

[1] RD53 Collaboration, "The RD53A Integrated Circuit," CERN-RD53-PUB-17-001 (2017).

[2] RD53 Collaboration, "RD53B Design Requirements," CERN-RD53-PUB-19-001 (2019).

[3] Xilinx, "Aurora 64B/66B Protocol Specification," SP011 (v1.3) October 1, 2014.

[4] M. Karagouins *et. al,* "An Integrated Shunt-LDO Regulator for Serial Powered Systems," in Proc. of IEEE ESSCIRC '09, (2009).

[5] A. ur Rehman *et. al,* "Performance simulations and characterization of RD53 pixel chips for ATLAS and CMS HL-LHC upgrades ," in TWEPP 2021 (2021) https://indico.cern.ch/event/1019078/contributions/4443947/.

[6] J. Christiansen, "RD53B users guide: Introduction to RD53B pixel chip architecture, features and recommendations for use in pixel detector systems." CERN-RD53-PUB-21-001 (2021)