High-sensitivity resonant cavity monitor system with direct RF sampling electronics[*]

Yi-Mei Zhou^{*, 1} Long-Wei Lai^{*, 1, †} Jia-Lan Pan,^{2, 3} Ying-Bing Yan,¹

2026-06-23

ChinaRxiv

Abstract

Resonant cavity probes are widely used in free-electron lasers and have been proposed for linear colliders to achieve high-resolution position and phase measurements. Electronics is a major limiting factor affecting system performance. For the C-band cavity beam arrival time measurement system used in the SHINE facility, a prototype cavity monitoring system employing direct RF sampling electronics was developed, operating at 3.52 GHz in the TM010 monopole mode. By replacing conventional down-conversion electronics, the direct RF sampling electronics simplifies the originally complex RF front end. The analog-to-digital converter used in the direct RF sampling processor has a -3 dB bandwidth of 9 GHz at 3.52 GHz, a maximum sampling rate of 2.6 GSPS, and an effective number of bits of 8. Theoretical analysis shows that the performance of the direct digital down-conversion electronics is comparable to that of an analog RF down-converter, and online beam tests at the Shanghai Soft X-ray Free-Electron Laser Facility confirmed its high sensitivity. At a charge of 80 pC, the electronics achieved a relative amplitude error of 1.3×10^{-4} and a relative phase error of 8.4 fs. Even at a bunch charge of 1 pC, the electronics maintained a relative amplitude error better than 1.0×10^{-3} and a relative phase error of 47.7 fs, which is beneficial for the large-scale application of direct RF sampling electronics in the SHINE facility. This represents a major advance in beam instrumentation for the SHINE facility, reducing project cost and improving system reliability.

Full Text

High-sensitivity resonant cavity monitor system with direct RF sampling electronics*

Yi-Mei Zhou, *1 Long-Wei Lai*, 1, † Jia-Lan Pan,2, 3 Ying-Bing Yan,1
Jian Chen,1 Xiao-Qing Liu,1 Shan-Shan Cao,1 and Ren-Xian Yuan1

1*Shanghai Advanced Research Institute, Chinese Academy of Sciences, Shanghai 201204, China*
2*Shanghai Institute of Applied Physics, Chinese Academy of Sciences, Shanghai 201800, China*
3*University of Chinese Academy of Sciences, Beijing 100049, China*

Resonant cavity monitors are widely used in free-electron lasers (FELs) and have been proposed for linear colliders to obtain high-resolution position and phase measurements. The read-out electronics is a major limiting factor affecting the system performance. A prototype cavity monitor system with direct radio frequency (RF) sampling electronics was developed for the C-band cavity beam arrival monitors used at SHINE, operating at the TM010 monopole mode frequency of 3.52 GHz. The direct RF sampling electronics simplified the otherwise complex RF front-end by applying traditional down-conversion electronics. The

direct RF sampling processor used analog-to-digital converters with −3 dB bandwidth of 9 GHz, maximum sampling rate of 2.6 GSPS, and 8 bits ENOB at 3.52 GHz. A theoretical analysis demonstrated that direct-digital down-conversion electronics performed comparably to analog RF down-converters, and online beam tests at SXFEL confirmed its high sensitivity. The relative amplitude error of the electronics was 1.3×10^{-4} at 80 pC, and the relative phase error was 8.4 fs. Even with a beam charge of 1 pC, the relative amplitude error of the electronics was better than 1.0×10^{-3}, and the relative phase error was 47.7 fs, facilitating large-scale applications of direct RF sampling electronics at SHINE. This represents a significant advancement in beam instrumentation at SHINE, which reduces project costs and enhances system reliability.

Keywords: SHINE, resonant cavity, direct RF sampling, electronics

I. INTRODUCTION

Accelerator-based advanced light sources are a cornerstone of modern scientific research. Among them, X-ray free-electron lasers (XFELs) stand out because of their ultrafast temporal resolution, ultrahigh spatial resolution, and extreme peak brightness [1–4], which make them indispensable for advancing research in fields such as life sciences, chemistry, physics, and materials science. Several XFEL facilities have been commissioned worldwide. China has progressed significantly in this domain, constructing high-gain free-electron laser facilities such as the Shanghai Deep Ultraviolet Free-Electron Laser (SDUV-FEL) [5–7], Dalian Coherent Light Source (DCLS) [8], and Shanghai Soft X-ray Free-Electron Laser (SXFEL) [9–13] and achieved multiple significant scientific breakthroughs during these projects.

Currently, the Shanghai High Repetition Rate X-ray FEL and Extreme Light Facility (SHINE) is under construction [14]. To date, SHINE, a continuous wave (CW) superconducting hard XFEL, is the largest scientific infrastructure project in China and is anticipated to drive transformative advancements in photon science upon completion. SHINE comprises a superconducting linear accelerator (LINAC) with a beam energy of 8 GeV, three undulator lines delivering photons in the range of 0.4 to 25 keV, three beamlines, and ten experimental end stations. The total length is 3.1 km, and the tunnel depth is 29 m underground [15, 16] (Fig. 1). SHINE was designed for a maximum bunch repetition rate of 1 MHz [17], with bunch intensities ranging from 10 pC to 300 pC. Currently, the injector section has been installed and commissioned, and SHINE is scheduled to be completed in 2027. The beam instrumentation system is designed to accurately measure and maintain the electron beam in XFELs [18], which is necessary for achieving high performance and efficient operation. The performance of this system has considerably evolved with the development of electronic technology. To achieve a high lasing efficiency in the undulators, high-resolution monitoring of the beam position, bunch arrival time, and intensity is required. Passive bunched-beam excited cavity resonators are widely used in XFELs owing to their high shunt impedance, which allows them to deliver output signals with a high signal-to-noise ratio (SNR). These resonators are commonly used for beam position monitoring (BPM) and bunch arrival monitoring (BAM) [19-23].

The dipole-mode frequency of the cavity output signal is primarily determined by the operating frequency of the RF gun, cutoff frequency of the beam pipe, and geometry and design of the cavity. Most cavity BPMs operate in the C-band (4 GHz -8 GHz), S-band (2 GHz -4 GHz) and X-band (8 GHz -12 GHz), while few operate in the L-band (1 GHz -2 GHz) and Ku-band (12 GHz -18 GHz). For example, for the Linac Coherent Light Source (LCLS-I, LCLS-II), a high-Q (Q: quality factor) cavity-BPM (CBPM) operating in the X-band was developed, which achieved a positional resolution of 200 nm at a charge of 200 pC [24]. A low-Q C-band CBPM developed in the European X-ray FEL (EXFEL), which achieved a positional resolution of 183 nm at a bunch charge of 180 pC and dynamic range of \$$\pm$\$500 μm [25]. The high-Q C-band CBPM developed for the SXFEL is expected to achieve a resolution of 177 nm at a bunch charge of 500 pC [26].

This work was supported by the National Natural Science Foundation of China (No.12175293), Youth Innovation Promotion Association, CAS (No.2019290), Outstanding Member of the Youth Innovation Promotion Association, CAS, and SHINE R&D and project

† *Corresponding author, lailw@sari.ac.cn;* These authors contributed equally to this work.

Fig. 1. (Color online) Schematic of the SHINE layout.

Read-out electronics are employed for RF signal processing in cavity monitors [27, 28]. High-sensitivity read-out electronics are essential for improving position resolution when extracting position or time (phase) information from the signals of a passive cavity resonator. However, the GHz frequency range of these signals prevents analog-to-digital converters (ADCs) belonging to previous generations from directly sampling these signals with the necessary resolution. Figure 2 illustrates the schematic layouts of three different RF signal-processing electronics [29]. Traditional methods implement direct analog I/Q demodulation (Fig. 2(a)) or utilize the heterodyne principle, down-conversion of the cavity BPM RF signals to an intermediate frequency (IF), followed by digital acquisition and demodulation (Fig. 2(b)). At the SACLA XFEL [22], EXFEL [30], and Swiss FEL [31] the cavity BPM RF signals are down-converted directly to the I/Q baseband, as shown in Fig. 2(a). This technique has the lowest requirements for ADCs and FPGAs; however, it faces the challenges of analog I/Q imbalance, local oscillator (LO) leakage, low-frequency noise interference, and twice the number of ADCs. In ATF [19], LCLS I/II [32, 33], SXFEL [26, 34] and several other FELs, the cavity BPM signals are down-converted to an IF (Fig. 2(b)), typically in the region of tens of MHz, and then digitized at a sufficiently high sampling rate to remain in the first Nyquist pass-band, typically 100–250 MSPS (mega samples per second). This structure reduces the difficulties in analog signal processing and achieves high gain by distributing the amplifier at different frequencies. Both structures contain long and complex analog signal processing chains and use complex local oscillator synthesizers, mixers, and multistage filters. Image rejection and local oscillator isolation must be considered; furthermore, analog RF components introduce noise and are sensitive to environmental changes. Figure 2(c) illustrates the direct RF sampling electronics, which digitizes the RF signal directly using an ADC and subsequently demodulates it through digital signal processing (DSP). DSP is advantageous over analog processing because it is programmable, free from drawbacks related to aging, tolerance, and environmental effects [35], and flexible, which enables the implementation of perfectly reproducible filters and algorithms for deterministically processing cavity RF signals. Although direct sampling has been successfully implemented for LLRF systems at 1.3 GHz in facilities such as the EXFEL [36] and Polish free electron laser (PolFEL) [37], extending this approach to C-band cavity monitors presents unique challenges. Our study demonstrates the first application of direct RF sampling at 3.52 GHz for cavity monitors, which requires (1) wider-bandwidth ADCs, (2) improved jitter performance, and (3) specialized signal processing to maintain precision at higher frequencies. At present, the sampling rates of ADCs that perform direct RF sampling exceeds 2 GSPS, with analog bandwidths of up to 10 GHz, and are commercially available with 12-bit or 14-bit resolution. This technological advancement facilitates the development of direct RF sampling electronics for SHINE.

Fig. 2. (Color online) Three types of cavity signal processing electronics.

Before this study, the feasibility of using direct RF sampling electronics for C-band cavity probe signal processing was experimentally verified in the laboratory using an RF signal generator, a direct RF sampling processor, and an amplification module designed specifically for the experiment [38]. These findings were utilized to design the cavity BPM/BAM system for SHINE incorporating direct RF sampling electronics presented in this paper. This study covered the design of

the cavity BPM/BAM, direct RF sampling electronics, and associated signal processing algorithms. The performance of the electronics was comprehensively analyzed and compared with that of a traditional IF down-conversion structure. In addition, online beam tests were conducted at the SXFEL to evaluate the system performance in a practical setting.

II. DESIGN OF CAVITY MONITORS FOR SHINE

Three types of cavity monitors are foreseen for SHINE: cavity BPMs for the distribution line, cavity BAMs for the LINAC, and cavity BPMs for the FEL sections. The cavity BPMs for the distribution line were designed with a 35 mm inner diameter beam pipe aperture, with a required resolution of 1 μm at 100 pC and dynamic range of $\pm$1 mm. The cavity BAMs for the LINAC, which also utilize a 35 mm beam-pipe aperture, require a resolution of 20 fs at 100 pC. For the cavity BPMs in the FEL sections, a smaller 8 mm beam-pipe aperture was used, with a required resolution of 200 nm at 100 pC and dynamic range of ± 100 μm [39]. This section presents the operational principles of cavity monitors and discusses the design considerations for the cavity BPM/BAM system at SHINE.

A. Cavity monitor principle

When a charged particle passes through a cylindrical cavity resonator, all eigenmodes with electric fields aligned along or near the beam axis are excited, which partly dissipates the beam energy into the resonator [40]. These excited modes carry different beam-related information, and their electromagnetic fields can be coupled to narrowband RF signals for beam observation. The TM010 monopole mode is proportional solely to the bunch charge, whereas the TM110 dipole mode is proportional to both the bunch charge and transverse position of the beam in the resonator.

A cylindrical cavity resonator serves as a beam charge monitor (BCM) and beam arrival time monitor (BAM) by extracting the TM010 mode signals [41, 42]. In addition, it functions as a beam position monitor (BPM) by utilizing the TM110 mode signals for the two polarizations aligned with the horizontal and vertical axes. Figure 3 illustrates the electromagnetic simulation of the two eigenmodes for a bunched electron beam passing through a cylindrical cavity.

In a cylindrical cavity BPM system [43], the TM110 mode is employed for beam position detection, where the transverse beam displacement induces signals (V_x, V_y) that are coupled out through the electrodes. Simultaneously, the TM010 mode provides a reference signal (V_r) proportional to the beam charge Q, thereby enabling normalization to mitigate charge fluctuation effects [44]. The relationships between the signals and beam parameters are given by

$$V_x \propto x \cdot Q,\ V_y \propto y \cdot Q,\ V_r \propto Q. \tag{1}$$

where x and y denote the horizontal and vertical beam displacements, respectively. The beam position is determined by

Fig. 3. (Color online) (a) TM010 and (b) TM110 modes in the cylindrical cavity.

normalizing the TM110 mode signal from the position cavity to the TM010 mode signal from the reference cavity as follows:

$$x = k_x * \frac{V_x}{V_r},\ y = k_y * \frac{V_y}{V_r}. \tag{2}$$

where k_x and k_y are sensitivity coefficients that can be determined through calibration experiments.

The voltage signal V_{out} from the cavity probe is modeled as a damped oscillation, described by

$$V_{\text{out}} = A_{\text{RF}} e^{-\frac{t}{2\tau}} \sin(\omega_{\text{res}} t + \phi). \tag{3}$$

Here, A_{RF} denotes the signal amplitude; $\omega_{\text{res}} = 2\pi f_{\text{res}}$, where f_{res} denotes the resonant frequency of the excited eigenmode; ϕ is the phase of the signal; and τ represents the decay time, which is related to the quality factor of the cavity.

The amplitude of the output signal A_{RF} depends on several factors, including the beam charge q, geometry of the cavity, and bunch length σ_z, is given by [20, 45]

$$A_{\text{RF}} = \frac{\omega_{\text{res}} q}{2} \sqrt{\frac{Z}{Q_{\text{ext}}} \frac{R}{Q} \exp\left(-\frac{\omega_{\text{res}}^2 \sigma_z^2}{2c^2}\right)}. \tag{4}$$

Here, Z denotes the characteristic impedance of the cavity, Q_{ext} the external quality factor describing the energy coupled to the readout port (P_{out}), and c the speed of light, respectively. Furthermore,

$$Q_{\text{ext}} = \frac{\omega_{\text{res}} U}{P_{\text{out}}}. \tag{5}$$

where U is the energy stored in the cavity, which dissipates via the cavity walls (P_{diss}). This depends on the wall material and is described by the following internal quality factor:

$$Q_0 = \frac{\omega_{\text{res}} U}{P_{\text{diss}}}. \tag{6}$$

The energy decay time τ is defined by the loaded quality factor Q_L, which accounts for both the external (Q_{ext}) and internal (Q_0) and is expressed as

$$\tau = \frac{Q_L}{\omega_{\text{res}}}. \tag{7}$$

where Q_{L} is given by

$$\frac{1}{Q_{\text{L}}} = \frac{1}{Q_0} + \frac{1}{Q_{\text{ext}}}. \tag{8}$$

The geometric shape factor of the cavity is R/Q (R-over-Q), which is primarily determined by the cavity structure. R/Q is calculated as follows:

$$\frac{R}{Q} = \frac{\left|\int E\, ds\right|^2}{\omega_{\mathrm{res}} U}. \tag{9}$$

where E and s denote the electric field along the beam axis and the path along the cavity, respectively. The electric-field distributions for the TM_{010} and TM_{110} modes are given by

$$E_Z = CJ_0\left(\frac{j_{01}r}{R}\right)e^{-i\omega_{010}t}. \tag{10}$$

$$E_Z = CJ_1\left(\frac{j_{11}r}{R}\right)\cos(\phi)e^{-i\omega_{110}t}. \tag{11}$$

where C is proportional to the amplitude of the oscillation; r is the radial coordinate; J_m is a Bessel function of the first kind of order m [46]; and j_{mn} is the nth root of the equation $J_m(j) = 0$. Here, $j_{01} = 2.405$, $j_{11} = 3.832$, and R denotes the cavity radius. The expressions for R/Q for both modes are given by

$$\left(\frac{R}{Q}\right)_{110} = \frac{2J_1^2\left(\frac{j_{11}r}{R}\right)LT^2}{\omega_{110}\varepsilon_0\pi R^2 J_0^2 j_{11}^2}. \tag{12}$$

$$\left(\frac{R}{Q}\right)_{010} = \frac{2LT^2}{\omega_{010}\varepsilon_0\pi R^2 J_1^2 j_{01}^2}. \tag{13}$$

where L is the cavity length; $T = \sin\left(\frac{\omega L}{2c}\right)\Big/\frac{\omega L}{2c}$ is the transit-time factor; and ε_0 is the vacuum dielectric constant.

B. Design of the cavity BPM and BAM for SHINE

To prevent interference and to ensure accurate measurements, the design frequency of cavity monitors must avoid the operating frequency of the RF cavities of SHINE, namely, 3.9 GHz [47, 48]. Therefore, to balance power efficiency, resolution, and manufacturing complexity, the cavity BPM monitors for SHINE were designed to operate in the C-band at frequencies of 3.52 GHz and 5.254 GHz, which were carefully selected to avoid overlap with the SHINE RF cavity frequency while providing high sensitivity for beam-position and arrival-time measurements. Furthermore, cavity BPMs were designed with high loaded quality factors (Q_{L}) to optimize performance and ensure stable and accurate signal detection. Table 1 summarizes the parameters of the two types of cavity BPM monitors planned for SHINE. Types 1 and 2 were designed for beam pipes with diameters of 35 mm and 8 mm [49], respectively. Both types were designed with a loaded Q_{L} that resulted in a decay time of 200 ns, which allowed the system to operate at a bunch-repetition rate of 1 MHz without aliasing and ensured reliable beam diagnostics over extended periods of operation.

Figure 4 illustrates the longitudinal cross-section and prototype of the cavity Beam Arrival Monitor (BAM) designed for SHINE. The key dimensions and parameters of the cavity are presented in Table 2. The measured value of the coupled output signal was approximately 0.5 V peak at $t = 0$ for a bunch charge of 100 pC.

Visible labels in Fig. 4(a): Feedthrough; Cavity; Beam Pipe; L_{gap}; R_{gap}; L_{cav}; D_{pipe}; R_{cav}; d_p.

Fig. 4. (Color online) (a) Structural design and (b) photograph of the SHINE CBAM.

A prototype of the cavity BAM for SHINE was fabricated and installed in the SXFEL for system evaluation. The SXFEL operated with a bunch charge of 500 pC, an RF frequency of 2856 MHz, and a synchronized clock of 119 MHz. Figure 5 shows the sampled raw data and spectrum of the BAM signal acquired using an oscilloscope with a 6 GHz bandwidth. The main spectral component of the signal is observed at 3.524 GHz, which is consistent with the design specifications of the BAM. An additional spectral compo-

Table 1. Parameters of the cavity BPMs at SHINE.

Cavity type	Type 1	Type 1	Type 2	Type 2
Parameters	Position	Reference	Position	Reference
Frequency /GHz	3.520	3.520	5.254	5.254
Radius /mm	51.9	32.6	34.8	21.8
Length /mm	9	4.5	9	5
Decay time /ns	200	200	200	200
Bandwidth /MHz	1.59	1.59	1.59	1.59
Loaded Q	2212	2212	3301	3301
Q_{ext}	3245	5869	5663	1268
Normalized shunt impedance	0.175@1mm	50.65	0.56@1mm	82.6
Sensitivity /V/nC	0.575@1mm	7.26	1.15@1mm	9.4

Table 2. Design of the cavity BAM in the SHINE.

Parameters	Value
Frequency /GHz	3.520
Decay time/ns	200
Bandwidth/MHz	1.59
Q_L	2213
Sensitivity /V/nC	5.2
D_{pipe} /mm	70

Parameters	Value
R_{gap} /mm	26
L_{cav} /mm	5
L_{gap} /mm	14.8
R_{cav} /mm	30
d_p /mm	1.7

nent at 4.36 GHz is observed, which arises from the cavity structure design. However, because the noise level of this secondary spectral component is approximately 20 dB lower than that of the main signal component, and subsequent signal processing is applied exclusively to the primary spectral component, the additional spectral component does not affect the measurement accuracy.

Fig. 5. (Color online) Output signal of the cavity BAM. (a) Time-domain waveform, (b) frequency spectrum.

III. DESIGN OF DIRECT RF SAMPLING ELECTRONICS PROTOTYPE

In contrast to traditional down-conversion systems, direct RF sampling electronics employ fewer analog components, thereby reducing the number of noise sources. However, this advantage must be carefully evaluated because ADCs in direct sampling systems typically operate over wider frequency ranges, integrate more noise bandwidth, and exacerbate jitter-induced noise. Although the reduced analog chain improves the SNR at lower frequencies, clock jitter significantly degrades the SNR of the ADC at higher input frequencies [50]. Consequently, beyond a certain frequency threshold, the overall SNR of a direct RF sampling system may fall below that of a down-conversion system. Additionally, high-performance direct-sampling ADCs and their precision clock references often incur higher costs, further complicating trade-offs in system design. The SNR tradeoff and its cost implications must be thoroughly analyzed to evaluate the suitability of direct sampling for practical applications.

In this section, we present the design of a direct RF sampling electronics prototype for the SHINE cavity BPM/BAM systems. For comparison, traditional down-conversion electronics were also developed. Performance analyses and online beam tests were conducted to evaluate and compare the performances of both electronic systems. These evaluations provide important insights into the advantages and limitations of the direct RF sampling approach for the beam-monitoring applications at SHINE.

A. Direct RF sampling prototype electronics

The two main components of the cavity monitor prototype electronics were a direct RF sampling processor and signal amplifier module.

1. *Direct RF sampling processor* A direct RF sampling processor was developed leveraging state-of-the-art electronics technologies, including RF direct sampling ADCs, high-performance system-on-chip (SoC) field-programmable gate arrays (FPGAs), and low-jitter phase-locked-loop (PLL) oscillators for clock signal generation. The processor was housed as a standalone unit within a 1U high, 19-inch enclosure. The system architecture comprised an FPGA carrier board, two FPGA mezzanine card (FMC) connector slots–one for an ADC board and the other for a timing board, based on the White Rabbit (WR) standard [51]. The WR timing system precisely synchronized the processor with other components in the SHINE system, facilitating accurate beam measurements and real-time data acquisition. This compact and modular design facilitated integration into the overall beam-monitoring infrastructure, while maintaining high performance and reliability.

The ADC board primarily consisted of two ADC chips (AD9689 from Analog Devices) and PLL circuits. Each ADC featured a bandwidth of 9 GHz, maximum sampling rate of 2.6 GSPS, and quantization resolution of 14 bits. The effective number of bits (ENOB) of the ADC were approximately 9 bits and 8 bits at input frequencies of 3.52 GHz and 5.25 GHz, respectively. The ADC chips utilized a dual-converter architecture with JESD204B-encoded serial digital outputs employing eight high-speed JESD204B lanes, each capable of achieving a data rate of up to 16 Gbps. Compared to CMOS and LVDS, the JESD204B serial data interface reduced the number of pins and minimized package size and conductor track complexity, which simplified the board design and reduced overall system costs. Thus, JESD204B is widely used in direct RF sampling ADC chips [52].

The sampling clock was synchronized with the external clock of the electron gun photocathode laser via a two-stage PLL system. This system comprised an LMX2582 PLL synthesizer and LMK04832 clock jitter cleaner (both from Texas Instruments). The PLL system ensured the synchronization of the sampling frequency over a broad range. For SHINE, the RF frequency was 1.3 GHz [53], and the external synchronized clock was set to 1.3/6 GHz, corresponding to 216.667 MHz. The ADC sampling clock can be synthesized to frequencies of up to 2.6 GHz. For the SXFEL with an RF frequency of 2.856 GHz, the external synchronized clock was 2.856/24 GHz, or 119 MHz, enabling the ADC sampling clock to be synthesized to 1.904 GHz [54, 55].

The carrier board for the direct RF sampling electronics was built around the Xilinx FPGA UltraScale+ multiprocessor system-on-chip (MPSoC) XCZU15EG, which integrated a 64-bit quad-core Arm® Cortex®-A53 processor, a dual-core Arm® Cortex-R5F processor, and Xilinx programmable logic (PL) into a single device. Additionally, it included on-chip memory, external multiport memory interfaces, and an extensive range of peripheral connectivity options [56]. The FPGA featured 24 gigabit transceiver high-speed (GTH) lanes that supported data rates of up to 12.5 Gbps and transmitted ADC data via the JESD204B protocol. Each ADC was supported by eight GTH lanes.

The integration of a processing system (PS) and programmable logic (PL) into a single chip significantly simplified the design of the digital board. The carrier board further incorporated four small form-factor pluggable (SFP) ports, dual data rate (DDR) memory for both PS and PL, interlock outputs, a user input/output system (IOS), a secure digital (SD) card slot, and a joint test action group (JTAG) interface. Figure 6 illustrates the block diagram and physical layout of the processor.

In addition, the processor was equipped with an FMC low-pin-count (LPC) slot that interfaced with the SHINE WR card, allowing the processor to receive a 1.003086 MHz trigger signal and bunch ID from the card for accurate data processing and synchronization. This setup ensures robust synchronization for system operation and efficient data handling.

Fig. 6. (Color online) Block diagram and photograph of the direct RF sampling processor.

2. ***RF signal amplifier*** Direct RF sampling electronics is primarily advantageous because it eliminates the need for analog down-conversion and streamlines the overall system architecture. However, the gain/attenuation stage remains essential for ensuring that the input signal to the ADC is optimal for accurate sampling. Therefore, an independent RF signal amplifier prototype was developed, which consisted of four channels housed within a single module. Figure 7 illustrates the structure of one channel, along with a photograph of the prototype.

Each RF amplifier channel consisted of a low-noise amplifier (LNA) PMA3-63GLN+ that provided a gain of up to 27.9 dB at 3.5 GHz, high-pass filter HFCN-2700, and digital-step attenuator RFSA3715 with an attenuation range of 0 to 31.75 dB. Although a high-pass filter was included in the pro-

prototype design, it was not strictly necessary for practical operation. The LNA amplified the RF signal with minimal noise contribution and conditioned the signal for optimal ADC input. The digital-step attenuator provided fine control over the signal amplitude, facilitating precise adjustments to maintain the signal levels required for ADC processing. This RF amplifier module was crucial for conditioning the RF signal and ensuring optimal performance of the direct RF sampling electronics within the SHINE cavity BPM/BAM system.

Fig. 7. (Color online) Schematic of one channel and a photograph of the prototype of the direct RF sampling processor.

B. RF down-conversion electronics for comparison

Along with the development of the direct RF sampling processor, a set of traditional down-conversion electronics was concurrently developed to enable performance comparison at the SXFEL. This system comprised an RF down-conversion module and a generic 1 GSPS beam signal processor for SHINE. The

RF down-conversion module included clock divider and multiplier circuits to generate a local oscillator (LO) signal at 3.466 GHz from the 2856 MHz input signal. It then down-converted the 3.52 GHz input signal to 54 MHz using a mixer and low-pass filter. In addition, the module incorporated bandpass filters, low-noise bypass amplifiers, and step attenuators to optimize signal quality.

Figure 8 shows the block diagram and a photograph of the front-end module and generic processor. A comparison with the direct RF sampling verification module (Fig. 7) shows that the down-conversion front-end module is considerably more complex and larger.

Fig. 8. (Color online) (a) Block diagram and photographs of the (b) RF down-conversion module and (c) generic processor for SHINE.

The design of the generic beam diagnostic processor developed for SHINE was similar to that of a direct RF sampling processor. It used the same carrier board as the direct RF sampling processor; however, AD9680 was used as the ADC board. The ADC board was equipped with four-channel ADCs, offering a 2 GHz bandwidth, a maximum sampling rate of 1.0 GSPS, and 14-bit quantization [39]. For operation in the SXFEL, the sampling rate was configured as 952 MSPS, which enabled optimal signal acquisition and processing for beam diagnostics at the facility.

C. Electronics performance analysis

The performance of the electronics is primarily determined by the analog components and ADC.

1. Analog components performance

The noise figure (NF) is a key parameter used to assess the degradation of the SNR as a signal passes through an electronic component or system. It is defined as the ratio of the SNR at the input to that at the output and is typically expressed in decibels (dB). The NF quantifies the amount of noise added by the device relative to an ideal noiseless system [57].

Mathematically, the NF is expressed as

$$NF = 10\log_{10}(F). \tag{14}$$

where F is the noise factor $F = \dfrac{S_i/N_i}{S_o/N_o}$; S_i/S_o is the power of the input/output signals of the gain stage; and N_i/N_o is the power of the input/output noise of the gain stage.

The NF of the cascaded gain stage G_n is calculated as follows:

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_4 - 1}{G_1 G_2 G_3} + \cdots. \tag{15}$$

where F_n and G_n are the NF and gain of the nth gain stage, respectively. The NF of the RF amplifier PMA3-63GLN+ at 3.5 GHz was 0.7 dB, and the gain was 27.9 dB. For passive components, such as the high-pass filter (HFCN-2700) and attenuator, the NF is equivalent to the insertion loss. The insertion loss of the high-pass filter HFCN-2700 was approximately 0.9 dB at 3.52 GHz, which resulted in an NF of 0.9 dB and a gain of −0.9 dB. The attenuator RFSA3715 had an insertion loss of approximately 2.2 dB at 3.52 GHz with an NF of 2.2 dB and a gain of −2.2 dB when it was set to 0 dB. The total NF of the RF signal amplifier module was 0.71 dB with a gain of 24.8 dB.

In comparison, the NF of the down-conversion RF module was calculated as 7.0 dB, with a gain of 22.0 dB. Consequently, the NF of the analog section of the direct RF sampling electronics was 6.3 dB lower than that of the down-conversion electronics, indicating significant improvement in the noise performance of the direct sampling approach.

The NF and gain were measured using Keysight' s FieldFox microwave analyzer (N9918B) at a frequency of 3.52 GHz. For the prototype amplitude-adjustment module of the direct RF sampling electronics, the NF was measured to be 2 dB, with a gain of 21 dB. In contrast, the down-conversion RF module exhibited an NF of 7.5 dB and a gain of 22.7 dB. Consequently, the NF of the analog section of the direct RF sampling electronics was 5.5 dB lower than that of the down-conversion electronics, demonstrating significantly improved noise performance.

2. *ADC performance*

The NF of the ADC is calculated as follows [58]:

$$NF = P_{\mathrm{FS}}(\mathrm{dBm}) + 174(\mathrm{dBm}) - \mathrm{SNR}_{\mathrm{ADC}} - 10\log_{10}\left(\frac{f_s}{2}\right). \tag{16}$$

where $P_{\mathrm{FS}}(\mathrm{dBm})$ is the full-scale power (dBm). The $P_{\mathrm{FS}}(\mathrm{dBm})$ values of both the direct RF-sampling ADC and 1 GSPS generic processors are identical. The SNRADC of the direct RF sampling ADC was approximately 58 dB at an input frequency of 3.52 GHz, whereas that of the 1 GSPS generic processor was approximately 67 dB at an input frequency of 54 MHz. Because the sampling rate f_s of the direct RF sampling ADC was twice that of the 1 GSPS ADC in SXFEL, the overall NF of the direct RF sampling ADC was 6 dB lower than that of the 1 GSPS ADC.

In comparison, the overall NF of the direct RF sampling electronics was nearly identical to that of the RF downconversion electronics. Theoretically, both electronics should offer equivalent performances when processing the signal from the cavity pickup.

3. *Digital signal processing in FPGA*

In the FPGA-based signal-processing system, digital data from the ADCs are used to extract the amplitude and phase information of the cavity signal. Owing to the narrow bandwidth of the cavity signal, several algorithms can be employed, such as digital down-conversion, Hilbert transform, and discrete Fourier transform (DFT). After considering factors such as the processing rate, resource utilization, and overall performance, the fixed-point Goertzel-DFT algorithm was selected for this application [38]. The Goertzel-DFT algorithm is suitable for applications where the signal frequency is predetermined and stable, as such as cavity monitoring systems. This algorithm efficiently computes the DFT at a specific frequency, minimizes the computational complexity, and optimizes FPGA resource usage.

$$X(k) = \sum_{n=0}^{N-1} x(n) e^{-j(\frac{2\pi}{N})kn}, \quad k = 0, 1, \dots, N-1. \tag{17}$$

where $x(n)$ denotes the digitized signal, N the number of sampled points, and K the extracted frequency, respectively. The results are complex values $X(k) = I(k) + j \cdot Q(k)$, where $I(k)$ and $Q(k)$ are the in-phase and quadrature-phase components, respectively. Subsequently, the signal amplitude V and phase ϕ can be extracted using the following equations:

$$V = \sqrt{I(k)^2 + Q(k)^2}. \tag{18}$$

$$\phi = \arctan\left(\frac{I(k)}{Q(k)}\right). \tag{19}$$

Figure 9 shows the waveforms and spectra of the sampled data obtained from the SHINE BAM prototype installed in the SXFEL. In total, 4096 points were sampled for each beam pulse. Owing to under-sampling, the center frequency of the cavity signal is shifted and converted to 285.9 MHz [59].

The relative amplitude error between the two channels of electronics can be calculated using the following expression:

$$R_v = \mathrm{std}\left(\frac{V_1 - V_2}{\left(\frac{V_1+V_2}{2}\right)}\right) / \sqrt{2}. \tag{20}$$

where V_1 and V_2 are the amplitudes calculated from the sampled data of the two channels.

The phase error of the electronics is defined as the standard deviation of the phase difference between the two channels:

$$R_p = \mathrm{std}(\phi_1 - \phi_2)/\sqrt{2}. \tag{21}$$

Fig. 9. (Color online) ADC data (a) waveform and (b) spectrum from direct RF sampling.

where ϕ_1 and ϕ_2 are the phase values calculated from the sampled data of the two channels.

The design bandwidth of the cavity signal was approximately 1.6 MHz. In the DFT algorithm, two methods can be used to obtain the peak amplitude of the cavity signal. First, the fixed-point DFT was calculated using all 4096 samples, where the frequency resolution of each point in the spectrum was 0.47 MHz. Subsequently, the points around the center frequency were summed to obtain a reasonable result. Figure 10(a) illustrates the amplitude obtained from the 4096-point DFT calculation, where sample 615 represents the peak value in the spectrum. Figure 10(b) shows the relative amplitude error as a function of the number of points accumulated around the peak point. The optimal result, with an error of 1.7×10^{-4}, is obtained by summing seven points. Figure 10(c) shows the phase error with the peak point exhibiting the best phase accuracy, resulting in a phase error of 14.9 fs. The phase error showed no further improvement when additional sample points were included in the phase calculation.

The second method involves selecting a subset of data from the starting point and calculating the DFT for its center frequency. Figure 11 shows the performance variations for different DFT lengths. Red points in Fig. 11(a) represent a 500-point subset of the complete acquisition. The best relative amplitude error, 1.7×10^{-4}, is achieved using 500–900 points, which is equivalent to the result from the first method. The phase error was 10.5 fs, which was slightly superior to that obtained using the first method.

Fig. 10. (Color online) (a) DFT amplitudes around the peak point, (b) relative amplitude error of different DFT accumulation points around the peak point, and (c) phase errors at different DFT points.

IV. BEAM TESTS AT THE SXFEL

The prototype cavity BAM for SHINE was manufactured and installed at the end of the LINAC section of the SXFEL

Fig. 11 panel text.

- (a) y-axis: ADC data; x-axis: Samples; scale: $\times 10^4$. Legend: "all sampled data" ; "intercepted processed data" .
- (b) y-axis: Relative Error (a.u.); x-axis: data lengths (points); scale: $\times 10^{-4}$.
- (c) y-axis: Phase Error (fs); x-axis: data lengths (points).

Fig. 11. (Color online) (a) Subset of sampled data, (b) relative amplitude error, and (c) phase error of the peak point DFT when computing different data lengths.

for online beam experiments and to evaluate the system performance. For the beam tests, the SXFEL operated at bunch intensities of approximately 500 pC. The output signal of the cavity BAM was transmitted from the tunnel to the cabinet via an approximately 40 meters long, 3/8 inch low-loss coaxial cable-type HELIAX LDF2-50. At the operating frequency of 3.52 GHz, the signal attenuation for the long cables is approximately 10 dB [60]. A four-channel power divider was used to split the signal; thus, the input signal to the electronics was equivalent to a bunch intensity of 80 pC.

A. The necessity of BPF

Band-pass filters are essential in RF analog heterodyne receivers because they reject image frequency mixing products and unwanted spurious sidebands and act as selective filters for the desired frequency range. Customarily, a BPF is positioned in front of the direct RF sampling electronics; however, the cavity resonator itself functions as an effective BPF owing to its narrowband band-pass filter characteristic. Furthermore, unlike down-conversion electronics, RF direct-sampling electronics do not generate spurious signals because of the direct sampling approach, in which no mixing processes contribute to spurious frequencies. In addition, the digital signal processing algorithm used in these systems extracts the signal at a fixed-point frequency, which further isolates the desired signal. Consequently, the BPF in direct RF sampling electronics for cavity-monitoring systems may be eliminated.

Figure 12(a) illustrates the test setup, where two 3.52 GHz cavity BPFs are connected in front of channels 1 and 2, whereas channels 3 and 4 are configured without BPFs. The relative amplitude error between channels 1 and 2 is 2.25×10^{-4}, whereas that for channels 3 and 4 is 1.29×10^{-4} (Fig. 12(b)). The phase RMS errors were 12.6 fs and 9.0 fs for channels 1 and 2 and for channels 3 and 4, respectively (Fig. 12(c)). These results indicate that the inclusion of the BPF does not improve the system performance of the SHINE cavity monitor; in fact, it slightly reduces the performance. Therefore, a BPF is not necessary in the configuration of the direct RF sampling read-out electronics of our system.

B. RF signal conditioning performance comparison

In Section III, the NF of the RF-signal amplifier shown in Fig. 7 is superior to that of the down-conversion electronics shown in Fig. 8(a). Figure 13 illustrates the test setup for comparing the performances of the two RF readout electronics. The four-way split cavity monitor signals were fed in pairs into the two channels of the two RF modules and then sampled using the direct RF sampling processor. Both signal paths were simultaneously measured using the same sampling processor (AD9689 ADC) under identical clock conditions. The performance differences reflect only the intrinsic characteristics of each signal conditioning approach. The relative amplitude errors for the direct RF and down-conversion modules are 1.67×10^{-4} and 2.63×10^{-4}, respectively, while the phase jitters are 9.6 fs and 11.0 fs, respectively. Although the SNR of the direct RF-sampling ADC was higher when sampling a down-converted 54 MHz IF input signal than when sampling a direct 3.52 GHz RF input signal, the performance of the RF conditioning prototype remained superior to that of the down-conversion RF front-end module.

Fig. 12. (Color online) (a) Test setup to evaluate the necessity of the BPF, (b) the amplitude correlation, and (c) phase errors with and without the BPF.

Fig. 13. (Color online) (a) Test setup to compare RF module performance, (b) the relative amplitude error, and (c) phase error.

C. High-sensitivity performance test

The charge of the SHINE beam bunches ranged from 10 pC to 300 pC. The electronics designed in this study were optimized for a bunch charge of 100 pC; however, the signal for the lower charge was too weak to be detected with a gain of 24.8 dB. Therefore, we cascaded two gain stages of the direct RF sampling module with a cable to double the power gain from 24.8 dB to 49.6 dB while maintaining the NF. However, this was not manageable in a simplistic manner for the analog RF down-conversion module. Attenuators with different values were added before the power splitter to simulate bunches with different charges. Figure 14 shows a block diagram of the test.

Fig. 14. (Color online) Block diagram of the test setup for system sensitivity.

Attenuators with different values are connected before the splitter, and the corresponding beam charges are listed in Ta-

Table 3 and Fig. 15. The lowest beam charge is 0.8 pC.

Table 3. (Color online) Relationship between the external attenuator and equivalent beam charge.

External Attenuator (dB)	Beam Charge (pC)
0	80
8	32
16	13
20	8
24	5
28	3
34	1.6
38	1
40	0.8

Fig. 15. (Color online) External attenuator and equivalent beam charge.

Figure 16 shows the relative amplitude and phase errors of the direct RF and down-conversion modules for different bunch charges. Both modules exhibit similar performance for bunch charges higher than 8 pC (with a 20 dB attenuator). At 80 pC (0 dB attenuator), the relative amplitude error of the direct RF sampling electronics is 1.3×10^{-4}, while that of the down-conversion electronics is 1.7×10^{-4}. As the charge continues to decrease, the amplitude and phase errors of the direct RF sampling electronics become smaller than those of the down-conversion electronics, owing to the higher gain of the direct RF sampling electronics. At a charge of 1 pC, the relative amplitude error is 9.57×10^{-4}, which is superior to the required value of 1.0×10^{-3} for SHINE. The charge and phase resolutions were 0.96 fC and 47.4 fs, respectively.

Fig. 16. (Color online) (a) Relative amplitude errors and (b) phase errors with different external attenuators.

V. CONCLUSIONS

A prototype cavity-monitoring system with direct RF sampling electronics was developed for SHINE and tested at the SXFEL. For the beam tests, a cavity monitor designated as BAM was used, which operated at a frequency of 3.52 GHz. Theoretical analysis showed that direct RF sampling electronics and down-conversion electronics exhibited approximately the same noise performance. For amplitude and phase extraction, a fixed-point DFT algorithm with a data length of 500–900 was the most effective. Online beam tests at the SXFEL confirmed that a BPF was not required. The relative amplitude and relative phase errors of the electronics were 1.3×10^{-4} at 80 pC and 8.4 fs, respectively. At lower charges, such as 1 pC, the relative amplitude error of the electronics remained below 1.0×10^{-3}, while the phase error was 47.7 fs. These results indicate that the cavity monitoring system with direct RF sampling electronics offers higher signal sensitivity than the traditional analog RF down-conversion technique when the RF gain stages are cascaded.

Based on the results of the above tests, new direct RF sampling electronics for the SHINE cavity BPMs is currently being developed. This represents the first large-scale application of direct RF sampling electronics for cavity monitors operating in the C band. The study represents a significant advancement in beam instrumentation at SHINE because direct RF sampling electronics play a crucial role in optimizing the system structure, enhancing performance, and reducing the cost of beam instrumentation in accelerators.

[1] Z.T. Zhao, C. Feng, X-ray free electron lasers. *Physics* **47**, 481–490 (2018). doi:10.7693/wl20180801

[2] N.S. Huang, H.X. Deng, B. Liu, Features and futures of X-ray free-electron lasers. *The Innovation* **2**, 100097 (2021). doi:10.1016/j.xinn.2021.100097

[3] Z.T. Zhao, C. Feng, K.Q. Zhang, Two-stage EEHG for coherent hard X-ray

generation based on a superconducting linac. *Nucl. Sci. Tech.* **28**, 117 (2017). doi:10.1007/s41365-017-0258-z

[4] C. Feng, H.X. Deng, Review of fully coherent free-electron lasers. *Nucl. Sci. Tech.* **29**, 160 (2018). doi:10.1007/s41365-018-0490-1

[5] Z.T. Zhao, Z.M. Dai, X.F. Zhao et al., The Shanghai high-gain harmonic generation DUV free-electron laser. *Nucl. Instrum. Meth. A* **528**, 591–594 (2004). doi:10.1016/j.nima.2004.04.108

[6] Z.T. Zhao, D. Wang, Seeded FEL experiments at the SDUV-FEL test facility. *IEEE Trans. Nucl. Sci.* **63**, 930–938 (2016). doi:10.1109/TNS.2016.2527678

[7] H.X. Deng, Z.M. Dai, On harmonic operation of Shanghai deep UV free electron laser. *Nucl. Sci. Tech.* **19**, 7–12 (2008). doi:10.1016/S1001-8042(08)60014-7

[8] H.L. Huang, Y. Yu, Y. Chang et al., Photodissociation dynamics of H2O at 111.5 nm by a vacuum ultraviolet free electron laser. *J. Chem. Phys.* **148**, 124301 (2018). doi:10.1063/1.5022108

[9] Z.T. Zhao, D. Wang, Q. Gu et al., Status of the SXFEL Facility. *Appl. Sci.* **7**, 607 (2017). doi:10.3390/app7060607

[10] Z.T. Zhao, D. Wang, Q. Gu et al., SXFEL: a soft X-ray free electron laser in China. *Synchrotron Radiat. News* **30**, 29–33 (2017). doi:10.1080/08940886.2017.1386997

[11] H.X. Deng, Z.M. Dai, Modified model for harmonic generation free electron laser. *Nucl. Sci. Tech.* **21**, 1–6 (2010). doi:10.13538/j.1001-8042/nst.21.1-6

[12] W.C. Fang, J.Q. Zhang, J.H. Tan et al., Design and preliminary test of the LLRF in C band high-gradient test facility for SXFEL. *Nucl. Sci. Tech.* **31**, 100 (2020). doi:10.1007/s41365-020-00806-6

[13] M.H. Song, C. Feng, D.Z. Huang et al., Wakefields studies for the SXFEL user facility. *Nucl. Sci. Tech.* **28**, 90 (2017). doi:10.1007/s41365-017-0242-7

[14] Z.Y. Zhu, Z.T. Zhao, D. Wang et al., SCLF: An 8-GeV CW SCRF linac-based X-ray FEL facility in Shanghai, in *FEL2017*, 38th International Free Electron Laser Conference, Santa Fe, NM, USA, Aug 2017, p. 182–184. doi:10.18429/JACoW-FEL2017-MOP055

[15] Z. Zhao, H. Ding, B. Liu et al., Shanghai hard x-ray FEL facility progress status, in *IPAC ' 24*, 15th International Particle Accelerator Conference, Nashville, TN, USA, May 2024, p. 967–972. doi:10.18429/JACoW-IPAC2024-TUZN1

[16] D. Gu, X. Li, Z. Wang et al., Physics Design and Beam Dynamics Optimization of the SHINE Accelerator, in *FLS2023*, 67th ICFA Advanced Beam Dynamics Workshop, Luzern, Switzerland, August 2023, p. 174–176. doi:10.18429/JACoW-FLS2023-WE4P13

[17] Y.F. Liu, R.P. Wang, J. Tong et al., Development of a high-repetition-rate lumped-inductance kicker magnet prototype for the beam switchyard of SHINE. *Nucl. Sci. Tech.* **35**, 37 (2024). doi:10.1007/s41365-024-01390-9

[18] V. Schlott, Free Electron Laser Beam Instrumentation and Diagnostics. *Synchrotron Light Sources and Free Electron Lasers*, 779–826 (2020).

[19] S. Walston, S. Boogert, C. Chung et al., Performance of a high resolution cavity beam position monitor system. *Nucl. Instrum. Meth. A* **578**, 1–22 (2007). doi:10.1016/j.nima.2007.04.162

[20] Y. Inoue, H. Hayano, Y. Honda et al., Development of a high-resolution cavity-beam position monitor. *Phys. Rev. Accel. Beams* **11**, 062801 (2008). doi:10.1103/PhysRevSTAB.11.062801

[21] R. Thorsten, F. Frommberger, W.C.A. Hillert et al., Measuring the intensity and position of a pA electron beam with resonant cavities. *Phys. Rev. Accel. Beams* **15**, 112801 (2012). doi:10.1103/PhysRevSTAB.15.112801

[22] H. Maesaka, H. Ego, S. Inoue et al., Sub-micron resolution rf cavity beam position monitor system at the SACLA XFEL facility. *Nucl. Instrum. Meth. A* **696**, 66–74 (2012). doi:10.1016/j.nima.2012.08.088

[23] C. Song, Y. Jia, Z. Liu et al., High-resolution cavity bunch arrival-time monitor for charges less than 1 pC. *Phys. Rev. Accel. Beams* **26**, 012803 (2023). doi:10.1103/PhysRevAccelBeams.26.012803

[24] A. Young, J. Dusatko, S. Hoobler et al., Performance Measurements of the New X-band Cavity BPM Receiver, in *IBIC ' 13*, 2nd International Beam Instrumentation Conference, Oxford, UK, September 2013, p. 735–738.

[25] M. Stadler, R. Baldinger, R. Ditter et al., Low-Q cavity BPM electronics for E-XFEL, FLASH-II and swissFEL, in *IBIC ' 14*, 3rd International Beam Instrumentation Conference, Monterey, CA, USA, September 2014, p. 670–674.

[26] J. Chen, Y.B. Leng, L.Y. Yu et al., Optimization of the cavity beam-position monitor system for the Shanghai soft X-ray free-electron laser user facility. *Nucl. Sci. Tech.* **33**, 124 (2022). doi:10.1007/s41365-022-01117-8

[27] B.P. Wang, Y.B. Leng, L.Y. Yu et al., Design and measurement of signal processing system for cavity beam position monitor. *Nucl. Sci. Tech.* **24**, 020101 (2013). doi:10.13538/j.1001-8042/nst.2013.02.007

[28] N. Zhang, B.P. Wang, Y.B. Leng, An on-line data acquisition system of oscilloscope-embedded input/output controller for cavity BPM measurement. *Nucl. Sci. Tech.* **22**, 321–325 (2011). doi:10.13538/j.1001-8042/nst.22.321-325

[29] W. Taylor, Considering GSPS ADCs in RF Systems. *Analog Devices*, November 2021. https://www.analog.com/en/resources/technical-articles/considering-gsps-adcs-in-rf-systems.html.

[30] M. Stadler, R. Baldinger, R. Ditter et al., Beam test results of undulator cavity BPM electronics for the European XFEL, in *IBIC' 12*, 1st International Beam Instrumentation Conference, Tsukuba, Japan, October 2012, p. 404–408.

[31] B. Keil, R. Baldinger, R. Ditter et al., First beam commission experience with the SwissFEL cavity BPM, in *IBIC' 17*, 6th International Beam Instrumentation Conference, Grand Rapids, MI, USA, August 2017, p. 251–254. doi:10.18429/JACoW-IBIC2017-TUPCF17

[32] S. Smith, S. Hoobler, R.G. Johnson et al., Commissioning and performance of LCLS cavity BPMs, in *PAC' 09*, 23rd Particle Accelerator Conference, Vancouver, BC, Canada, May 2009, p. 754–756.

[33] J. Frisch, P. Emma, A. Fisher et al., Electron beam diagnostics and feedback for the LCLS-II, in *FEL' 14*, 36th International Free Electron Laser Conference, Basel, Switzerland, August 2014, p. 666–671.

[34] J. Chen, Y.B. Leng, L.Y. Yu et al., Beam test results of high Q CBPM prototype for SXFEL. Nucl. Sci. Tech. **28**, 51 (2017). doi: 10.1007/s41365-017-0195-x

[35] M. Wendt., Direct (Under) sample vs. analog down conversion for BPM electronics. in *IBIC' 14*, 3rd International Beam Instrumentation Conference, Monterey, CA, USA, September 2014. p. 486–494.

[36] Z. Geng, S. N. Simrock, Evaluation of fast ADCs for direct sampling RF field detection for the European XFEL and ILC. in *LINAC08*, 24th Linear Accelerator Conference, Victoria, BC, Canada, September 2008. p. 1030–1032.

[37] T. Kowalski, G. P. Gibiino, J. Szewinski et al., Experimental Evaluation of Sub-Sampling IQ Detection for Low-Level RF Control in Particle Accelerator Systems. Sensors. **22**, 38 (2022). doi: 10.3390/s22010038

[38] R. Meng, L. Lai, Y. Zhou et al., RF direct sampling and processing electronics for SHINE cavity BPM system. Nucl. Instrum. Meth. A. **1063**, 169256 (2024).doi: 10.1016/j.nima.2024.169256

[39] Y.X. Han, L.W. Lai, Y.M. Zhou et al., High performance generic beam diagnostic signal processor for SHINE. in *IBIC' 24*, 13th International Beam Instrumentation Conference, Beijing, China, September 2024. p. 385–388. doi:10.18429/JACoW-IBIC2024-WEP53

[40] P.B. Wilson, J.E. Griffin, High energy electron linacs; application to storage ring RF systems and linear colliders, AIP conference proceedings, 1982, p. 450–555. doi:10.1063/1.33620

[41] S.S. Cao, Y.B. Leng, R.X. Yuan et al., Optimization of beam arrival and flight time measurement system based on cavity monitors at the SXFEL. IEEE Trans. Nucl. Sci. **68**, 2–8 (2021). doi: 10.1109/TNS.2020.3034337

[42] C.Y. Song, W.H. Huang, Noninvasive beam diagnosis based on the TM010 mode. Nucl. Sci. Tech. **35**, 127 (2024). doi: 10.1007/s41365-024-01409-1

[43] J. Chen, Y.B. Leng, L.Y. Yu et al., Study of the crosstalk evaluation for cavity BPM. Nucl. Sci. Tech. **29**, 83 (2018). doi: 10.1007/s41365-018-0418-9

[44] S. Zorzettis, M. Wendt, H. Mainaud-Durand, Characterization and alignment of a resonant cavity beam position monitor for the Compact Linear Collider (CLIC) project at CERN. Nucl. Instrum. Meth. A. **1048**, 167898 (2023). doi: 10.1016/j.nima.2022.167898

[45] T. Nakamura, Development of beam-position monitors with high position resolution. Master' s thesis, The University of Tokyo, 2008.

[46] C.F. du Toit, The numerical computation of Bessel functions of the first and second kind for integer orders and complex arguments. IEEE Trans. Antenn. Propag. **38**, 1341–1349 (1990). doi: 10.1109/8.56985

[47] Y.X Zhang, J.F. Chen, D. Wang et al., RF design optimization for the SHINE 3.9 GHz cavity. Nucl. Sci. Tech. **31**, 73 (2020). doi: 10.1007/s41365-020-00772-z

[48] Y.M Zhen, J.F. Chen, Design optimization of 3.9 GHz fundamental power coupler for the SHINE project. Nucl. Sci. Tech. **32**, 132 (2021). doi: 10.1007/s41365-021-00959-y

[49] J. Chen, S.S. Cao, R.X. Yuan et al., Development and performance evaluation of the CBPM system for the SHINE. in *IBIC ' 24*, 13th International Beam Instrumentation Conference, Beijing, China, September 2024.

[50] Z.J. Towfic, S.K. Ting, A.H. Sayed, Clock Jitter compensation in high-rate ADC circuits. IEEE Trans. Signal Process. **60**, 11 (2012). doi: 10.1109/TSP.2012.2208958.

[51] Y.B. Yan, P.X. Yu, G.H. Chen et al., White rabbit based beam-synchronous timing systems for SHINE. in *IPAC2022*, 13th International Particle Accelerator Conference, Bangkok, Thailand, June, 2022. doi:10.18429/JACoW-IPAC2022-THIYGD1.

[52] J. Harris, What is JESD204 and why should we pay attention to it? Analog Devices, August 2019. https://www.analog.com/en/technical-articles/what-is-jesd204-and-why-should-we-pay-attention-to-it.html.

[53] Y.M Zhen, S.J. Zhao, X.M. Liu et al., High RF power tests of the first 1.3 GHz fundamental power coupler prototypes for the SHINE project. Nucl. Sci. Tech. **33**, 10 (2022). doi: 10.1007/s41365-022-00984-5

[54] J.G. Wang, B. Liu, Development of readout electronics for bunch arrival-time monitor system at SXFEL. Nucl. Sci. Tech. **30**, 82 (2019). doi: 10.1007/s41365-019-0594-2

[55] J.G. Wang, X.Q. Liu, L. Feng et al., Analysis of electro-optical intensity modulator for bunch arrival-time monitor at SXFEL. Nucl. Sci. Tech. **30**, 3 (2019). doi: 10.1007/s41365-018-0524-8

[56] Zynq UltraScale+ MPSoC data sheet: overview. AMD Xilinx, November, 2022 https://www.amd.com/content/dam/xilinx/support/documents/data_sheets/ds891-zynq-ultrascale-plus-overview.pdf.

[57] I. Rosu, Understanding noise figure. https://www.qsl.net/va3iul/Noise/Understanding%20Noise%20Figure

[58] W. Kester, ADC noise figure - an often misunderstood and misinterpreted specification. Analog Devices. https://www.analog.com/media/en/training-seminars/tutorials/MT-006.pdf.

[59] D. Osterland, A. Benzin, F. Gerfers et al., Undersampling and SNR Degradation in Practical Direct RF Sampling Systems. 2023 European Conference on Networks and Communications & 6G Summit (EuCNC/6G Summit): Physical Layer and Fundamentals (PHY), June 2023.doi: https://doi.org/10.1109/EuCNC/6GSummit58263.2023.10188304

[60] LDF2-50 datasheet. CommScope. https://www.commscope.com/globalassets/digizuite/115591-p360-ldf2-50-external.pdf.

Note: Figure translations are in progress. See original paper for figures.