

Kelvin-Sensed AC Wheatstone Bridge Front End with Coherent Complex Ratiometric Readout for FLASH Radiotherapy Graphite Calorimetry

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1. Introduction

FLASH radiotherapy delivers a therapeutic dose within a few hundred milliseconds at average dose rates above 40 Gy/s. Under these conditions the ionization chambers used for conventional reference dosimetry suffer severe volume recombination, and their correction factors become strongly dependent on dose per pulse. An independent detector whose response does not rest on any ion-collection assumption is therefore required in order to establish traceable reference dosimetry at ultra-high dose rates.

Graphite calorimetry meets this requirement because absorbed dose follows directly from a measured temperature rise and a known specific heat capacity. The difficulty is entirely one of instrumentation rather than of principle. One gray deposited in graphite raises the core temperature by only 1.41 mK, and the whole irradiation is over in roughly 0.1 s, so the readout must resolve a sub-millikelvin step within a bandwidth of a few hertz.

Reported FLASH and ultra-high-dose-rate calorimeters read the thermistor bridge with a direct-current excitation and a nanovoltmeter or a digital multimeter. That architecture places the measurement band at direct current, where the amplifier 1/f noise, the thermoelectric offsets at connector junctions and the slow drift of the bridge arms are all at their largest. It also ties the sampling rate to the integration period of the meter, which is hard to reconcile with a delivery lasting a tenth of a second.

This work moves the measurement band away from direct current by exciting the bridge with a 1024 Hz carrier and recovering the envelope by software lock-in detection. Three features distinguish the resulting front end from a conventional bridge amplifier. The bridge nodes are sensed through a four-terminal Kelvin connection so that cable resistance drift does not enter the measurement. The excitation is digitized simultaneously with the bridge output at an exact integer multiple of the carrier frequency. The two demodulated quantities are then combined as a complex ratio, which removes excitation amplitude and phase drift from the result.

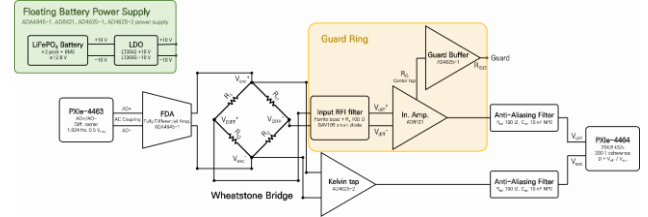


Fig. 1. Schematic of Kelvin-Sensed AC Wheatstone Bridge

2. Bridge Topology and Required Sensitivity

2.1 Opposite-arm configuration

The calorimeter core carries two NTC thermistors of nominal 10 kΩ at 25 °C with $\beta = 3892$ K (Amphenol Thermometrics, St. Marys, PA, USA), embedded symmetrically about the core center. They occupy opposite arms of a full Wheatstone bridge whose remaining two arms are ultra-low-drift metal-foil resistors of the same nominal value.

This arrangement was chosen over a quarter bridge for two reasons. Both thermistors respond to the radiation-induced temperature rise with the same sign of bridge imbalance, so the differential output doubles while the two uncorrelated thermistor noise sources add in quadrature. The signal-to-noise ratio therefore improves by a factor of two rather than by $\sqrt{2}$. In addition, a linear temperature gradient across the core produces equal and opposite resistance changes at the two symmetric positions and is largely rejected, which relaxes the requirement on core thermal uniformity and on the symmetry of the surrounding aerogel insulation.

2.2 Sensitivity

The relative resistance sensitivity of the thermistor is $\alpha = \beta/T^2 = 4.38$ %/K at 25 °C, so a 1 Gy dose gives $\Delta R/R = 6.18 \times 10^{-5}$. For an excitation of 0.5 V and the opposite-arm topology the differential bridge output is

$$V_{out} \approx (V_{exc}/2) \cdot (\Delta R/R) = 15.4 \mu V \text{ per Gy.}$$

This figure sets the target for the front end. The amplifier must have an input-referred noise small compared with the thermal noise of a 10 kΩ source, and a gain that places the signal comfortably inside the input

range of the digitizer without saturating on residual bridge offset.

2.3 Excitation level and self-heating

The excitation amplitude is bounded from above by thermistor self-heating. At 0.5 V the current through each branch is 25 μA and the power dissipated in each thermistor is 6.25 μW . For a dissipation constant of the order of 1 mW/K in the assembled probe this corresponds to a steady temperature elevation near 6 mK, which is several times the temperature rise produced by a 1 Gy delivery.

The elevation is not an error, since it is static, appears equally in both thermistors and cancels in the balanced bridge. What matters is that the sensitivity grows linearly with the excitation while the self-heating grows quadratically. Doubling the excitation to 1 V would raise the sensitivity to 30.9 $\mu\text{V}/\text{Gy}$ but would raise the self-heating to about 25 mK, which begins to shift the thermistor operating point and to disturb the thermal equilibrium between the core and the surrounding jacket. An excitation of 0.5 V was adopted as the compromise between these two trends.

3. Analog Front-End Design

3.1 Excitation and differential drive

The 1024 Hz carrier is generated by a PXIe-4463 source (National Instruments, Austin, TX, USA) and applied to the board single-ended. An ADA4945-1 fully differential amplifier (Analog Devices, Wilmington, MA, USA) converts it to a balanced drive for the bridge. The balanced drive was preferred to single-ended excitation because it holds the bridge common-mode potential at a level set on-board through the output common-mode pin, independent of the source ground. Common-mode current injected through the calorimeter cable therefore does not modulate the bridge balance.

3.2 Kelvin sensing of the bridge nodes

The calorimeter head is connected to the board through a cable of several meters carrying the excitation current, so sensing the bridge nodes on those same conductors would place the cable resistance in series with the bridge arms. The absolute value of that resistance is not the problem, since a static imbalance can be nulled. The problem is its temperature coefficient. Copper has a resistance temperature coefficient of about 0.39 %/K, so a lead resistance of 1 Ω in series with a 10 k Ω arm produces a relative change of 3.9×10^{-7} per kelvin of cable temperature. Compared with the 6.18×10^{-5} per gray of the thermistors, this is equivalent to 6.3 mGy for every kelvin of temperature change along the cable. A treatment room whose temperature drifts by one

or two kelvin during a session would generate an apparent dose of the order of ten milligray, which exceeds the noise floor of the front end by an order of magnitude.

Each bridge node is therefore brought out on a separate pair of conductors, one carrying the excitation current and one sensing the potential. The sense conductors terminate at the high-impedance inputs of the instrumentation amplifier, so they carry essentially no current and their resistance and its drift produce no voltage error. The four-terminal connection is carried across the board connector and up to the amplifier input pins, since a shared return trace on the board would reintroduce the same error in miniature.

3.3 Instrumentation amplifier stage

The sensed bridge output is amplified by an AD8421 instrumentation amplifier (Analog Devices, Wilmington, MA, USA) at a gain of 100. Its input voltage noise density of 3.2 nV/ $\sqrt{\text{Hz}}$ is well below the 12.8 nV/ $\sqrt{\text{Hz}}$ Johnson noise of the 10 k Ω bridge source impedance, so the measurement remains source-limited rather than amplifier-limited. Its common-mode rejection ratio exceeds 120 dB at the carrier frequency, which is required because the differential signal of interest rides on a common-mode carrier four orders of magnitude larger.

Input protection is provided by BAV199 low-leakage clamp diodes (Nexperia, Nijmegen, the Netherlands) and series resistors. Leakage current, not clamp voltage, is the governing specification here, because a leakage of a few nanoamperes into a 10 k Ω arm produces a static imbalance comparable to several gray, which makes a general-purpose clamp diode unusable. An ADA4625 stage (Analog Devices, Wilmington, MA, USA) drives the input guard traces at the input common-mode potential and buffers the excitation monitor, so that surface leakage across the board does not appear as a differential error and does not defeat the Kelvin connection described above.

3.4 Anti-alias filtering

A passive RC low-pass filter follows the gain stage. It is deliberately kept passive and of low order, since an active filter would add its own noise and offset drift within the signal band while the delta-sigma converters of the PXIe-4464 already provide sharp digital anti-alias filtering. The passive stage only has to attenuate energy near the modulator sampling rate and limit slew from any radiation-induced transient at the amplifier output. Keeping the stage passive also keeps its phase response stable with temperature, which matters because the phase of the demodulated signal is used in Section 4.

3.5 Power and grounding architecture

The board accepts a floating dual supply of ± 12.8 V and regenerates ± 10 V locally with an LT3042 and an LT3093 ultra-low-noise regulator pair (Analog Devices, Wilmington, MA, USA). Local regeneration makes the supply noise seen by the front end a property of the on-board regulators rather than of the cable and the external supply. Because the input is floating, the only path that defines the bridge common-mode potential is the deliberate on-board reference, which removes the ground loop that would otherwise couple mains-frequency current into the bridge.

Two details of the regulator configuration proved important. The power-good feedback divider was removed and the feedback pin tied directly to the input, because the divider presented a leakage path that shifted the output by an amount comparable to the noise target. The reference set capacitor was increased from 10 nF to 0.22 μ F, which lowers the reference noise corner into the sub-hertz region and keeps the regulator contribution negligible at the carrier frequency. A C0G dielectric was specified so that the capacitance does not vary with bias or temperature.

4. Coherent Ratiometric Readout

4.1 Coherent simultaneous sampling

The amplified bridge output and the excitation monitor are digitized on two channels of a PXIe-4464 digitizer (National Instruments, Austin, TX, USA) running at 204.8 kS/s. This rate is exactly 200 times the 1024 Hz carrier frequency, and both the source module and the input module derive their clocks from the same PXIe backplane reference.

Two consequences follow. First, any demodulation window whose length is a whole number of samples contains a whole number of carrier cycles, so the discrete Fourier transform of the record has no spectral leakage at the carrier bin and no window function is needed. Leakage would otherwise place a fraction of the large carrier amplitude into the neighboring bins that carry the demodulated dose signal, and since the carrier is four orders of magnitude larger than the signal, even a small leakage fraction would dominate the measurement. Second, because the two channels are sampled simultaneously rather than multiplexed, the phase relation between the excitation and the bridge output is preserved sample by sample and is not corrupted by an inter-channel skew that would drift with temperature.

Demodulation also places a component at 2048 Hz alongside the baseband term, far from the band of interest, so the low-pass stage of the demodulator removes it without any compromise to the 5 Hz signal bandwidth.

4.2 Complex ratiometric normalization

Both digitized channels are demodulated against the same reference to give in-phase and quadrature

components, so the bridge output becomes a complex quantity and the excitation becomes a second complex quantity. The measurement is then formed as the complex ratio of the two rather than as the bridge output alone.

Any variation common to both paths cancels in this ratio. Source amplitude drift, cable attenuation and phase shift, driver gain drift and residual regulator ripple all affect the numerator and the denominator in the same proportion, so the ratio depends only on the bridge imbalance. This is stronger than the usual ratiometric argument for direct-current bridges, because the complex form cancels phase as well as amplitude.

The imaginary part is not discarded. A purely resistive imbalance, which is what a dose deposition produces, appears entirely in the real part, while cable capacitance imbalance, dielectric absorption and leakage paths appear in the imaginary part, where they can be monitored to verify that the measurement remains resistive. The excitation monitor buffer is therefore treated as a signal path with the same layout discipline as the bridge path rather than as a diagnostic output.

5. Noise Budget and Expected Performance

Table I summarizes the input-referred noise contributions at the carrier frequency.

Table I. Input-referred noise budget

Contribution	nV/ $\sqrt{\text{Hz}}$
Bridge Johnson noise (10 k Ω)	12.8
AD8421 voltage noise	3.2
ADA4945-1, referred to input	<1
Total (quadrature sum)	13.2

The front end is therefore source-limited, as intended. Combining 13.2 nV/ $\sqrt{\text{Hz}}$ with the 15.4 μ V/Gy sensitivity gives the dose resolutions of Table II.

Table II. Projected dose resolution against noise bandwidth

ENBW (Hz)	Noise (nV)	Dose (mGy)
1	13.2	0.9
5	29.6	1.9
25	66.1	4.3

A 5 Hz bandwidth corresponds to a settling time of the order of 0.1 s, which matches the duration of a FLASH delivery, so the design should resolve individual deliveries of a few gray with a signal-to-noise ratio of several hundred.

Placing the measurement band at 1024 Hz is what makes these figures achievable. The band sits far above the $1/f$ corner of the instrumentation amplifier and above the mains harmonics, so the densities in Table I are white-noise densities rather than the much larger low-frequency values that a direct-current bridge would see. The Kelvin connection of Section 3.2 removes the cable drift term, and the complex ratio of Section 4.2 removes

the excitation drift term, so the residual systematic error is expected to be dominated by the stability of the metal-foil reference arms and by the thermal equilibrium of the calorimeter head rather than by the electronics.

6. Conclusions

A preamplifier board has been designed to read a two-thermistor graphite calorimeter bridge with a 1024 Hz carrier for FLASH radiotherapy dosimetry. The opposite-arm bridge gives 15.4 $\mu\text{V}/\text{Gy}$ at 0.5 V excitation, an excitation level fixed by the trade-off between a sensitivity that grows linearly and a self-heating that grows quadratically. A four-terminal Kelvin connection eliminates a cable drift term worth 6.3 mGy per kelvin, coherent sampling at exactly 200 samples per carrier cycle suppresses spectral leakage from a carrier four orders of magnitude larger than the signal, and complex ratiometric normalization against the simultaneously digitized excitation cancels amplitude and phase drift common to both paths. The resulting input-referred noise of 13.2 nV/ $\sqrt{\text{Hz}}$ is dominated by the Johnson noise of the bridge itself and corresponds to a projected dose resolution near 2 mGy at a 5 Hz bandwidth.

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