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M. TANAKA, H. IKEDA, K. TAMAI,
M. TAKEMOTO and H. HAYASHII

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Development and Evaluation of Charge-Sensitive Preamplifier for CsI Calorimeter in the KEK B-factory

Manobu Tanaka, Hirokazu Ikeda, Kunio Tamai,
National Laboratory for High Energy Physics, KEK
1-1 Oho, Tsukuba, 305 Japan
Megumi Takemoto *and Hisashi Hayashii
Nara Women's University
Kita-uoya nishi-machi, Nara, 630 Japan

Abstract

We developed a charge-sensitive preamplifier for the electromagnetic calorimeter (CsI(Tl)) for the KEK B-factory. We examined not only noise characteristic and dynamic range but also the stability of the preamplifier. In order to study the stability of the preamplifier, the input/output impedance and the power-supply ripple-rejection ratio were measured for various loads at the input/output node.

1 Introduction

In the KEK B-factory the barrel part of the electromagnetic calorimeter comprises 7000 CsI(Tl) crystals. Each crystal has two photodiodes (HAMAMATSU S2744-03) and two preamplifiers. The capacitance of the photodiode is 100 pF at a bias voltage of 60 V and the leakage current is less than 4 nA. Since a photodiode-readout has several advantages as a photon sensor over phototubes [1], a photodiode has been used for CsI(Tl) calorimeters[2],[3],[4] and for BGO,[5],[6], where charge-sensitive preamplifiers were commonly employed. In the B-factory experiment we measure the energy of photons and electrons over an energy range between 10 MeV and 8 GeV[7].

At the low-energy region (i.e. below 0.1 GeV), in order to maintain an energy resolution of better than $1.3\%/\sqrt{E}$ we must maintain an equivalent noise charge below 700 electrons for a detector capacitance of 100 pF[7]. In order to achieve this requirement several JFETs were examined for transconductance and $1/f$ noise using a Quateck Semiconductor Noise Analyser. Furthermore, we confirmed the noise characteristics of selected JFETs by a test circuit board of a preamplifier. Details were described in [8]. Since we had to mount two channels in a space of $5 \times 5 \times 2 \text{ cm}^3$, it was desirable that the input JFET for the preamplifier was a mini-mold type package. We tested 2SK322, 2SK613, 2SK1576, 2SK980 and 2SK932. The JFETs 2SK322 and 2SK613 showed a similar good performance. Though 2SK322 have been used for TRISTAN experiment[9], we chose 2SK613 for the input device because it was difficult to obtain a stable supply of 2SK322.

Since we chose the feedback capacitance of 1 pF for the preamplifier, the output voltage of the preamplifier was about 3 V for an incident energy of 8 GeV. We chose the decay time of 50 μsec for the preamplifier. The dynamic range of the preamplifier was specified as 10 V to be large enough for our application.

We have to consider the stability of the preamplifier circuit. Real parts of the input and output impedance must keep positive for any loading conditions. Since ripples on power rails are main reason of external noise, it is desirable that the power-supply ripple-rejection ratio is high.

We describe a circuit configuration of the preamplifier in section 2 and the performance in section 3. We summarized this paper in section 4.

2 Circuit configuration

Figure 1 shows the circuit configuration of the preamplifier. V_{CC} and V_{EE} are 15 V and -5 V. The charge gain is 1 V/pC, and the power consumption is 120 mW. The circuit is realized by a hybrid circuit on a four-layered printed-circuit board. We chose to use parallel-plate capacitors for C1 and C2 using small area on the hybrid circuit. A corner frequency of Q1, at which the device-intrinsic white-noise and the $1/f$ noise contributed equally to the noise, was 1 kHz.

While the preamplifier's topology is similar to the previous work published in [10], we put some new features to our circuit. When ripples are fed into power rails of the preamplifier, the ripples are transferred by Q4, Q6, Q7 and Q8 to OUT-E and OUT-C. In order to suppress the effect we have to achieve a high power-supply ripple-rejection ratio (PSRR). In traditional way a passive filter have been used for filtering on power rails of a preamplifier. If we used a resistance of small value and a chip-type capacitor, it would be difficult to achieve a low cutoff frequency of the filter (i.e. lower than a few kHz.). We added filters on V_{CC} and V_{EE} , which comprised Q5, C5, C7, R8 and Q9, C6, C8, R7. If current gains of Q5 and Q9 is large, the cutoff frequency of the filter is determined by time constants of C7, R8 and R7, C8. Since we set the cutoff frequency of 16 Hz, the PSRR became high.

*present address Mitsubishi Co. Ltd.

Figure 1. Circuit configuration of the preamplifier.

The output stage of the preamplifier is a Darlington combination of Q2 and Q10. We chose a JFET as the device of Q2 in order to avoid a degradation of the open loop gain for large signal output. The preamplifier has OUT-C and OUT-E so as to drive a transmission line from either the emitter node or the collector node of the preamplifier. We are going to use OUT-C as a signal output-node.

Figure 2. Time constant dependence of the equivalent noise charge. The solid lines are results of the fitting.

3 Performance

3.1 Noise

We measured the equivalent noise charge (ENC) for various detector capacitances (C_d) and various shaping times using a pulse generator (HP8110A), a shaper (ORTEC 450 Research Amplifier) and an oscilloscope (Tektronix

11403A). A test pulse was fed into TEST, and we added C_d at IN. OUT-C was pulled up by 1 k Ω to V_{CC} . The output signal from OUT-C was terminated at the input of the shaper by 50 Ω after an AC-coupling capacitor. The shaping function of the shaper was described in [11]. We chose the integration time and the differentiation time of the shaper to be equal. Figure 2 shows ENC as a function of the shaping time constant. The ENC is optimized at the time constant of around 1.5 μ sec for $C_d=100$ pF. Since the ENC is less than 400 electrons at the detector capacitance of around 100 pF, the preamplifier satisfies our requirement for ENC.

In order to understand the behaviour of ENC we fitted the measurement values by a theoretical equation [12]. Since a contribution of $1/f$ noise of about 36 electrons at $C_d=100$ pF for $CR(RC)^2$ shaping, we considered the part of the $1/f$ noise for the fitting. The fitting parameters were stray capacitance (C_s), the transconductance (g_m) and the leakage current (i_{leak}) at the IN node. We assumed the junction capacitance of Q1 was 6.6 pF from a data sheet. The filtering function of the shaper was approximated to a Poisson filter ($CR(RC)^2$). An error from the approximation was $\pm 2\%$. The fitting results were $C_s=(30 \pm 6.8)$ pF, $g_m=(20 \pm 2.5)$ mS and $i_{leak}=(0.0 \pm 1.9 \times 10^{-2})$ nA. We measured the stray capacitance of the test bench and the printed-circuit board by a C-V analyzer (KEITHLEY 590). Transistors connected IN node (Q1 and Q3) were removed from the printed-circuit board when we measured the stray capacitance. The stray capacitance was 15 pF on the test bench, and 10 pF on the printed-circuit board of the preamplifier. The stray capacitance on the printed-circuit board contributes to the total detector capacitance by 10 %. We could reduce the stray capacitance by changing the layout pattern on the circuit board.

3.2 Dynamic range

We measured the integral nonlinearity at OUT-E node and the OUT-C node by an oscilloscope (Tektronix 11403A). A test pulse was fed into TEST of the preamplifier. The biasing condition of the preamplifier was the same as that for the noise measurement. The preamplifier drove a $50\ \Omega$ transmission cable, and the cable was terminated at the input of the oscilloscope. Figure 3 shows the integral nonlinearity of the preamplifier at OUT-E and OUT-C. The output voltage is 5 V at OUT-E for 10 V test pulse because of the back termination of R2. The output voltage is 10 V at OUT-C for an input test pulse of 10 V. Both integral nonlinearities are less than 0.3% for the test pulse amplitude of 10 V.

Figure 3. Integral nonlinearity.

3.3 Impedance

The input and output impedances were measured by a network analyzer (ADVANTEST R4611) together with a SWR bridge. Figure 4 shows the real part of the input impedance as a function of the frequency.

Figure 4. Frequency dependence of the input impedance.

The input impedance was measured at $R2=8.6$ and $51\ \Omega$. The real part of the input impedance was positive in the full scale range of the measurement system (i.e. 10

MHz $\sim$ 300 MHz.). The SWR bridge limited the lower side of the range, and the upper limit came from the network analyzer. Figure 5 shows the real part of the output impedance. We measured the output impedance for $C_d=0$, 51 and 100 pF. The real part of the output impedance was maintained larger than $51\ \Omega$. We confirmed stable operation of the preamplifier with a photodiode of $C_d=100$ pF.

Figure 5. Frequency dependence of the output impedance. The solid line, the dashed line and the dotted line are guide to the eyes.

3.4 PSRR

We measured PSRRs for V_{CC} and V_{EE} at OUT-C and OUT-E. Figure 6 shows the setup when we measured PSRR at OUT-C for V_{CC} . When we measured PSRR at OUT-E, OUT-C node was terminated by $51\ \Omega$ via a capacitor of $50\ \mu\text{F}$. V_{EE} was supplied from the output node of LH0032 when we measured PSRR for V_{EE} . The input power from the network analyser was -10 dBm. Figures 7 (a) and (b) show PSRR of V_{CC} as a function of the frequency and that of V_{EE} . We added C_d of 100 pF at IN in the measurement. PSRR for V_{CC} and for V_{EE} was less than -60 dB around 1 MHz.

Figure 6. Test-bench set up for a PSRR measurement for V_{CC} .

Figure 8 shows PSRR as a function of detector capacitance. We measured PSRR at 2 kHz and 1 MHz for $C_d=0$, 10, 51, 100 and 200 pF. PSRR at 2 kHz was saturated

above 51 pF, and there was no capacitance dependence of PSRR at 1 MHz. We can achieve a PSRR of -40 dB at 2 kHz and -60 dB around 1 MHz. This high PSRR is enough for our application.

4 Summary

A charge-sensitive preamplifier was developed for the CsI(Tl) calorimeter. The preamplifier was examined for noise, dynamic range, input impedance, output impedance, and power supply ripple rejection ratio. The equivalent noise charge was less than 400 electrons for a detector capacitance of 100 pF. The optimum time constant of the shaper (ORTEC 450) was about 1.5 μ sec. The integral nonlinearity of the preamplifier was less than 0.3% for maximum output voltage of 10 V. We measured the input and output impedances of the preamplifier. The preamplifier worked stably for a detector capacitance of 100 pF. We measured the power supply ripple rejection ratio for V_{CC} and V_{EE} . We were able to achieve a PSRR of -40 dB at 2 kHz and a PSRR of less than -60 dB around 1 MHz.

Figure 7. Frequency dependence of PSRR. (a) shows PSRR for V_{CC} . (b) shows PSRR for V_{EE} .

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Figure 8. Detector capacitance dependence of the PSRR for V_{CC} .

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