The DuAmp - An Alternative Input Configuration for Detector Preamplifiers

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Abstract

An alternative type of detector-preamplifier configuration (DuAmp) has been assembled and tested* . The circuit employs two identical charge sensitive preamplifiers with the detector connected between the two preamplifier inputs. One of the preamplifiers is powered by voltage supplies referenced to ground potential, while the second one is powered by isolated supplies referenced to the bias voltage of the detector. The detector is biased by the voltage difference between the inputs of the two preamplifiers. The output of one of the preamplifiers is inverted and added to the output signal of the other preamplifier. The resultant signal is processed by pulse shaping electronics. For our application, the circuit has been shown to exhibit an overall noise level that is lower than that obtained using an equivalent single preamplifier in a conventional manner. A preliminary analysis of the circuit shows that the improvements are to be expected only in circumstances in which the ratio of detector to preamplifier capacitance is small.

I. INTRODUCTION

Traditionally, high resolution detector systems employ a charge sensitive preamplifier connected either directly or through a decoupling capacitor to the detector [1]. The output of the preamplifier represents the charge pulse from the detector together with the noise associated with the detector and the preamplifier itself.

The optimization of low-noise amplifiers has been analyzed in great detail and design procedures are well developed [2,3]. These techniques can be applied in a straightforward manner if one has full freedom to tailor input devices and circuitry to a given application. In many instances, however, practical necessity may dictate use of an existing preamplifier design with a given detector, so the question arises whether an alternative connection scheme may yield performance superior to the usual configuration. This paper describes such an alternative method of improving the performance of non-optimum systems. It should be made clear at the outset that this scheme does not improve on the noise achievable in an optimized system. Indeed, simple design modifications will often yield superior results, with both lower power and component count. This paper is directed specifically toward users who need to make the best of existing amplifiers, although the theoretical analysis may be of interest to circuit designers as well.

We have investigated an alternative configuration that utilizes two amplifiers sensing the signal current at opposite sides of the detector. This configuration also allows application of the detector bias voltage through the second amplifier, rather than through a separate high-value bias resistor with its attendant noise and parasitic capacitance. We call this configuration a DuAmp.

II. CIRCUIT CONFIGURATION

Fig. 1 shows the DuAmp block diagram. The detector **D** is connected between the inputs of two identical charge sensitive preamplifiers **PA1** and **PA2**. While **PA2** is powered by voltage supplies referenced to ground potential, **PA1** is powered by isolated supplies referenced to the bias voltage $\mathbf{V_b}$ of the detector. Due to the feedback of the preamplifiers, the

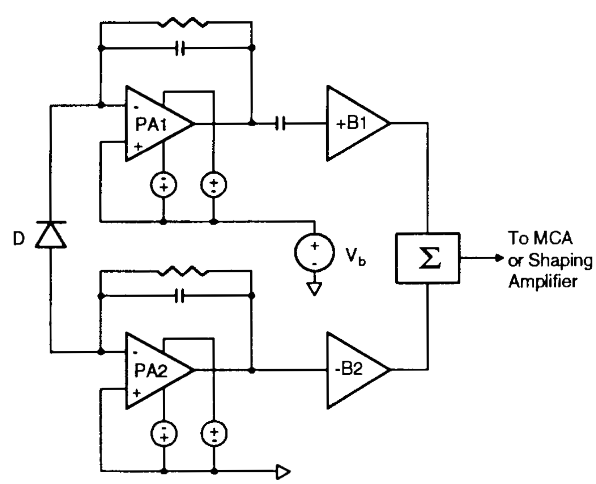

Fig. 1 DuAmp detector-preamplifier configuration.

* It was our initial impression that the concept of the DuAmp configuration was original. In subsequent conversations with F. Goulding and D. Landis, we have learned that similar or identical concepts were included in unpublished discussions during technical meetings held early in the development of nuclear pulse processing.

voltage difference between the inputs of **PA1** and **PA2** is equal to $\mathbf{V_b}$. Thus, the detector **D** is biased by the voltage difference between the inputs of the two preamplifiers.

The charge produced by the detector (signal, leakage current) is sensed by both amplifiers. The output signals from **PA1** and **PA2** have the same magnitude but are of opposite polarity. After inverting the signal from **PA2**, both signals are added together in a summation unit **Σ**. The inverter **-B2** following **PA2** and the buffer **+B1** following **PA1** may have additional functions of pole-zero cancellation and pulse shaping. If the signals from the preamplifiers are shaped before the summation, the output of **Σ** can be applied directly to the input of a multichannel analyzer. Otherwise, the summation is followed by pole-zero cancellation and pulse shaping.

The assembly of this configuration was originally motivated by the thought that some improvement in signal-to-noise (S/N) characteristics could be expected based on the following argument: the noise sources inherent to the detector result in correlated noise at the outputs of the two preamplifiers, while the contribution of preamplifier noise is uncorrelated between the two branches. This potential improvement is mitigated by the presence of feedback across each amplifier and their coupling through the detector capacitance. Our experimental observations confirm a net improvement for the applications described in this paper in which the detector capacitance is relatively small.

III. CIRCUIT ANALYSIS

Consider an amplifier that, with a detector capacitance C_d at the input, yields a noise voltage at the output v_n. For a given detector signal the signal-to-noise ratio at the amplifier output will be V_s/v_n. Now consider a pair of amplifiers with the same characteristics connected to the same detector as shown in Fig. 2. From the point of view of each amplifier the opposite detector electrode is effectively grounded through the low input impedance of the second amplifier, so the shunt capacitance at the input is the same as in the single amplifier configuration. The detector signals at the amplifier outputs are $V_{s1} = V_s$ and $V_{s2} = -V_s$, and the output noise levels are $v_{n1} = v_{n2} = v_n$. In the absence of any cross-coupling of noise between the two amplifiers, subtracting the two amplifier outputs from one another yields a total signal $V_{s,tot} = V_{s1} - V_{s2} = 2V_s$ and a total noise $v_{n,tot} = \sqrt{v_{n1}^2 + v_{n2}^2} = v_n\sqrt{2}$. The resulting signal-to-noise ratio is $2V_s / v_n\sqrt{2}$, which is $\sqrt{2}$ greater than in the original single-amplifier configuration.

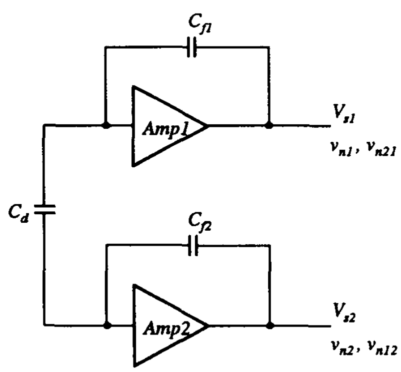

Fig. 2 Idealized dual amplifier circuit model.

In reality, however, noise from amplifier 1 is coupled to amplifier 2 and vice versa. This can be analyzed quite easily in the idealized configuration shown in Fig. 2. Amplifier 1 can be viewed as an inverting voltage amplifier, with an input signal v_{n2} applied through the impedance Z_{i1} formed by the series connection of C_{f2} and C_d to the input of amplifier 1, and $Z_{f1} = C_{f1}$ completing the feedback network. The voltage gain of this configuration (since no current can flow into the infinite input impedance of amplifier 2) is

$$A_{vx} = -\frac{Z_{f1}}{Z_{i1}} = -C_{f1} \cdot \left(\frac{1}{C_{f2}} + \frac{1}{C_d}\right) \tag{1}$$

provided the open loop gain of the amplifier >> Z_{f1}/Z_{i1}. If, as is usual in a good design, $C_d >> C_{f2}$

$$A_{vx} \approx -\frac{C_{f1}}{C_{f2}} = -1 \tag{2}$$

with the consequence that amplifier 2 introduces an additional noise contribution at the output of amplifier 1

$$v_{n21} = -v_{n2} \tag{3}$$

By the same argument, amplifier 1 introduces an additional noise component at the output of amplifier 2

$$v_{n12} = -v_{n1} \tag{4}$$

In adding up the noise contributions we must keep in mind that while v_{n1} and v_{n2} are not correlated, the cross-coupled contributions are anticorrelated. Since contributions due to v_{n1} (and, correspondingly, v_{n2}) are fully correlated, they add algebraically, so that when the two amplifier outputs are subtracted the correlated noise components are

$$v_{n1\Sigma} = v_{n1} - v_{n12} = 2v_{n1} \tag{5}$$

and

$$v_{n2\Sigma} = v_{n2} - v_{n21} = 2v_{n2} \tag{6}$$

Since $v_{n1\Sigma}$ and $v_{n2\Sigma}$ are not correlated they add in quadrature, yielding the total noise

$$v_{n,tot} = \sqrt{(2v_{n1})^2 + (2v_{n2})^2} = 2 \cdot \sqrt{v_{n1}^2 + v_{n2}^2}. \qquad (7)$$

If $v_{n1} = v_{n2} = v_n$

$$v_{n,tot} = 2\sqrt{2} \cdot v_n \qquad (8)$$

so that in this configuration the overall signal-to-noise ratio $2V_s / (2\sqrt{2} \cdot v_n)$ is inferior to that of the single amplifier by $1/\sqrt{2}$.

Practical amplifiers differ in an important respect from the configuration shown in Fig. 2. There is always a shunt capacitance C_p from each amplifier input to ground, as indicated in Fig. 3. This includes the capacitance of the input

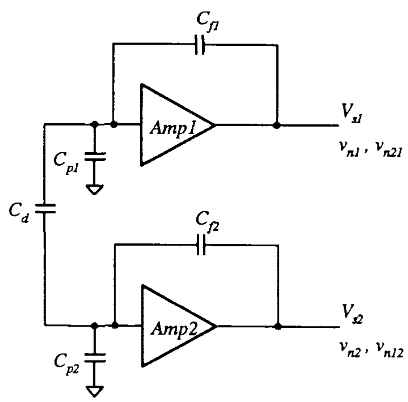

Fig. 3 Practical dual amplifier circuit model.

transistor and stray capacitance of the input connections. In calculating the cross-coupled noise in this configuration, the capacitances C_{f1}, C_{f2}, C_{p1}, C_{p2}, C_d will also be represented by their impedances Z_{f1}, Z_{f2}, Z_{p1}, Z_{p2}, Z_d (where $Z = 1/\omega C$). First, consider the cross-coupling of v_{n2} to the output of amplifier 1. Assuming that the impedances Z_d, $Z_{p2} << Z_{f2}$, the current flowing through C_{f2} is

$$i_{f2} = \frac{v_{n2}}{Z_{f2}}. \qquad (9)$$

At the input of amplifier 2 the current divides between the shunt capacitance C_{p2} and the detector capacitance C_d, so that the current flowing through C_d into the input of amplifier 1 is

$$i_{i1} = i_{f2} \frac{C_d}{C_d + C_{p2}}. \qquad (10)$$

Since the input of amplifier 1 is a virtual ground, the impedance determining the current flow to the amplifier is determined by C_d alone. Current flowing into the input node of amplifier 1 can flow into both the shunt capacitance C_{p1} and the feedback capacitor C_{f1}. For simplification we assume that the open loop gain A_{v0} of the amplifier is large enough that $A_{v0}C_{f1} >> C_{p1}$ so practically all of current i_{i1} flows through C_{f1}. Then the resulting voltage at the output of amplifier 1 is

$$\begin{aligned} v_{n21} &= i_{i1} \cdot Z_{f1} = - \frac{v_{n2}}{Z_{f2}} \frac{C_d}{C_d + C_{p2}} \cdot Z_{f1} = \\ &= - v_{n2} \cdot \frac{Z_{f1}}{Z_{f2}} \frac{C_d}{C_d + C_{p2}} \end{aligned} \qquad (11)$$

or, since $Z_{f1} = Z_{f2}$

$$v_{n21} = - v_{n2} \cdot \frac{C_d}{C_d + C_{p2}} \qquad (12)$$

Although this expression includes approximations that do not always apply in practical systems, it demonstrates the basic phenomena. The cross-coupled noise contribution depends on the ratio of detector capacitance to shunt capacitance C_p at the amplifier input. If $C_d >> C_p$, the cross-coupled noise is maximal, and the overall signal-to-noise ratio is degraded by $\sqrt{2}$ as shown for the simplified configuration in Fig. 2. The case $C_d << C_p$ appears attractive because of the $\sqrt{2}$ improvement in S/N, but it is not a desirable one, since the noise determining capacitance is dominated by the amplifier (or strays) rather than by the detector, so one could easily do better with a better amplifier design. In the intermediate regime, where the detector capacitance is comparable, but smaller than C_p, the two-amplifier readout can provide a noticeable advantage if an optimized amplifier is not available. Specifically, under the condition

$$\frac{C_d}{C_d + C_p} < \sqrt{2} - 1 \qquad (13)$$

or

$$\frac{C_d}{C_p} < \frac{1}{\sqrt{2}} \qquad (14)$$

the dual amplifier readout yields lower noise than the single amplifier.

At given detector capacitance C_d the noise levels $v_{n1} = v_{n2} = v_n$ increase with increasing shunt capacitances $C_{p1} = C_{p2} = C_p$. At the same time, however, the coupled noise components $v_{n12} = v_{n21}$ decrease with increasing C_p. Assuming that the current fluctuations in the channel of the input FET are the dominant noise source, the signal-to-noise ratio in case of single amplifier configuration is given by

$$(SN_s)^2 = \left(\frac{S}{v_n}\right)^2 = k \cdot \frac{C_p}{(C_p + C_d)^2} \qquad (15)$$

where k is a constant. $(SN_s)^2$ has a maximum at $C_p = C_d$. Thus, the maximum signal-to-noise ratio in case of single amplifier is

$$(SN_s)^2_{max} = k \cdot \frac{1}{4C_d} \qquad (16)$$

In the case of the dual amplifier configuration the total noise can be obtained from equations (5), (6) and (12)

$$v_{n\Sigma}^2 = 2 \cdot v_n^2 \cdot \left(1 + \frac{C_d}{C_d + C_p}\right)^2 = 2 \cdot v_n^2 \cdot \frac{(2C_d + C_p)^2}{(C_d + C_p)^2} \qquad (17)$$

The signal-to-noise ratio $(SN_D)^2$ is then given by

$$(SN_D)^2 = \left(\frac{2S}{v_{n\Sigma}}\right)^2 = \left(\frac{S}{v_n}\right)^2 \cdot \frac{2 \cdot (C_d + C_p)^2}{(2C_d + C_p)^2}. \qquad (18)$$

After combining equations (15) and (18), $(SN_D)^2$ can be expressed as

$$(SN_D)^2 = k \cdot \frac{2C_p}{(C_p + 2C_d)^2}. \qquad (19)$$

A maximum in signal-to-noise ratio in the case of dual amplifiers is achieved when $C_p = 2C_d$. This maximum value is equal to the maximum signal-to-noise ratio obtained in case of an optimized single amplifier

$$(SN_D)^2_{max} = k \cdot \frac{1}{4C_d}. \qquad (20)$$

This analysis shows that the predicted signal-to-noise ratio attainable with an optimized DuAmp configuration is the same as that from an optimized single preamplifier. However, the optimum occurs when the input capacitance of the preamplifiers is twice the value of the detector capacitance, whereas the optimum for a single preamplifier is reached when the capacitances are matched. DuAmp makes better use of amplifiers whose input capacitance is larger than the detector capacitance and would be non-optimum if used singly.

IV. First Experimental Results

A DuAmp circuit has been built and tested using two model A250 preamplifiers manufactured by Amptek Inc. [4]. A250 is a low noise, hybrid charge-sensitive preamplifier designed for use with wide variety of detectors. Its compact size and low operating power make it a suitable choice for the DuAmp configuration. Both preamplifiers were powered by two sets of alkaline batteries each providing ±6V supply voltage. The circuit was tested using a Hamamatsu S1223 PIN photodiode operating as X-ray detector at room temperature [4]. The PIN photodiode biasing voltage was set to 25V (V_b in Fig. 1).

Both preamplifiers were connected to separate ORTEC 575 shaping amplifiers . The amplifier in the "biased" branch of DuAmp (the upper branch in Fig. 1) was set to non-inverting mode while the other was operated in inverting mode. The shaping constants of both amplifiers were set to 1.5 μs. The output signals were summed in an inverting amplifier and, after another inversion, the summed signal was sent to a multichannel analyzer.

For comparison purposes, the energy resolution of the detector was first measured using a single preamplifier in conventional mode. Fig. 4 shows an Am-241 spectrum obtained with the single preamplifier configuration. The energy resolution (FWHM) for the well defined peaks at 59.54 keV and 13.95 keV was found to be 1.66 keV and 1.57 keV respectively.

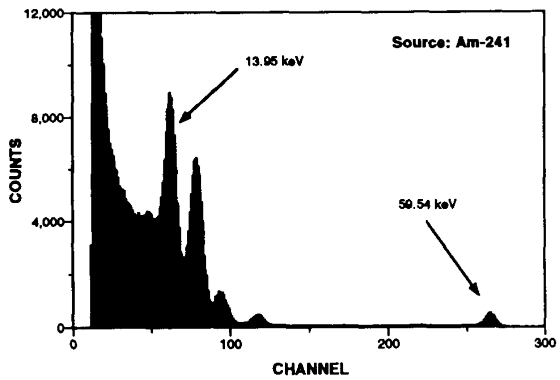

Fig. 4 Am-241 spectrum obtained with a single preamplifier configuration.

Substituting the DuAmp configuration with two identical A250 units, an energy resolution of 1.52 keV and 1.40 keV was measured for the same energy peaks, corresponding to an improvement of 140 eV and 170 eV respectively. An example of an Am-241 spectrum obtained using the DuAmp is shown in Fig. 5.

A second test was made using a photodiode with lower noise characteristics and A250 preamplifiers with a surface-mount input FETs. In this test, the separate shaping

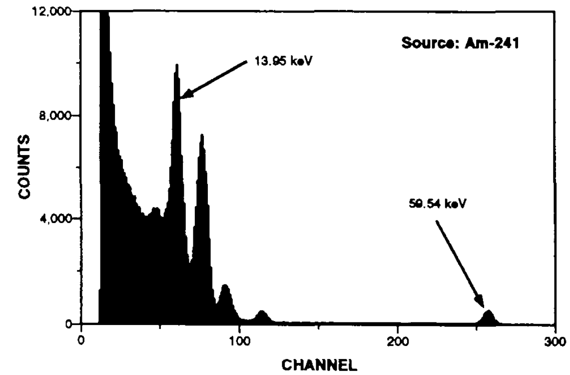

Fig. 5 Am-241 spectrum obtained with the DuAmp configuration.

amplifiers (**B1** and **B2** in Fig. 1) were replaced with a single ORTEC 450 research amplifier with differential input to minimize the possible influence of any noise or pickup beyond the preamplifiers. The energy resolution at 13.95 keV was observed to be 1.02 and 1.06 keV for each preamplifier used in conventional single mode. When reconfigured in the DuAmp mode, the energy resolution improved to 0.96 keV.

Further tests with these components were made by measuring the energy resolution under different configurations. As one test, the coupling effect was observed by measuring the deterioration of the energy resolution in one branch in single mode due to the presence of the second branch. With the second branch connected and used to supply bias voltage but without signal summing, the energy resolution worsened from 1.02 to 1.06 keV in one branch, and from 1.06 to 1.09 keV in the other.

The effect of detector capacitance was checked by placing a dummy capacitance of 5 pF across the detector terminals. Under those circumstances, the resolution in DuAmp mode became slightly worse (1.12 keV) compared with single mode operation (1.11 keV).

Further checks were carried out using the SPICE model to verify the effect of capacitance values on the noise behavior. These simulations show that the DuAmp configuration results in better performance for low detector capacitance, but becomes inferior to single mode operation at higher capacitance values compared with the preamplifier shunt capacitance.

V. CONCLUSION

The DuAmp configuration has resulted in an improved signal-to-noise performance for our application involving a detector with small capacitance. In general, however, there may be no advantage over a conventional single preamplifier if it is specifically matched to the detector of interest. The applications of the technique may therefore be limited to those circumstances in which the preamplifier can not be conveniently tailored to the specific detector employed in a given application. This may be the case, for example, when using commercially-available preamplifiers with shunt capacitance larger than the detector capacitance.

VI. ACKNOWLEDGMENTS

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VII. REFERENCES

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