Broadband Signal Amplification Paths for Semiconductor Radiation and Particle Detectors (Review)

E. A. Puryga at al.

Abstract—This study is devoted to the development of amplifying paths for plasma diagnostics, a significant part of which use semiconductor detectors as sensors that form low-intensity current signals. One feature of these diagnostics is the oscillographic form of registering sensor signals. In tandem with detectors, broadband transimpedance amplifiers that are based on operational amplifiers are used to amplify and normalize sensor signals. The principles of constructing such amplifying paths are considered taking the factors affecting their final noise and frequency characteristics into account. Practical examples of the construction of amplifying paths of corpuscular and neutron plasma diagnostics, as well as the Thomson-scattering diagnostics, that are used at the plasma installations of the Institute of Nuclear Physics (Siberian Branch, Russian Academy of Sciences) are presented.

INTRODUCTION

When conducting research in the field of high-temperature plasma physics and controlled thermonuclear fusion (CTF), diagnostics are widely used that are focused on registering the spatial distributions of the plasma parameters and their temporal dynamics, on the formation of feedback signals that stabilize the plasma-filament position in a magnetic trap, the formation of the necessary radial profiles of its density and temperature, as well as the intensity of fusion reactions. Such diagnostics include microwave, optical, X-ray, corpuscular, and other diagnostics; the key feature of their measuring paths is associated with the need to record sensor signals in an oscillographic form. The actually existing structure of an elementary measuring path includes a sensor, a broadband amplifier, a high-speed analog-to-digital converter (ADC), a digital unit based on a user-programmable gate matrix or a processor that generates measurement results using real-time procedures for digital signal processing, as well as an interface unit that transmits the results to the diagnostic server and/or the corresponding controller. A significant part of the measuring paths that are used in research on plasma physics and CTF, semiconductor diodes are used as detectors that form a low-intensity broadband current signal under the influence of radiation or particles entering their aperture. Such signals are amplified and normalized traditionally using transimpedance amplifiers (TIAs) based on low-noise broadband operational amplifiers (OAs) [1]. The diagram of such a path amplifier is shown in Fig. 1.

CONSTRUCTION OF AN AMPLIFICATION PATH BASED ON A TRANSIMPEDEANCE AMPLIFIER

If the characteristics of the basic OA are ideal, the output voltage of the TIA is determined by the values of the feedback resistance and photodiode current:

$$e_0 = i_{\text{sig}} K_\infty = i_{\text{sig}} R_f,$$

where $i_{\text{sig}}$ is the current of the semiconductor detector and $R_f$ is the resistance in the feedback circuit (Fig. 1).

For real OAs, this formula is valid only for the low- and medium-frequency regions, in which their gain with the open feedback $A(f)$ is large and actually constant. As the frequency increases, $A(f)$ decreases:

$$A(f) = \frac{A_\infty}{1 + \frac{f}{f_c}},$$

where $f_c$ is the amplifier cutoff frequency. As a consequence, an additional element, the error factor $\beta A(f)$, appears in formula (1) [2]:

$$\frac{\beta A(f)}{1 + \beta A(f)}.$$
where $\beta$ is the feedback coefficient of the amplifier. Figure 2a presents the amplitude–frequency (frequency response) and phase–frequency characteristics of the amplifier (FR and PFC) at the zero input capacitance. In this case, the feedback coefficient $\beta$ is represented by a constant, while the loop gain $A(f)$ decreases at a rate of 20 dB/decade with increasing frequency, as does the FR of the amplifier. The maximum phase shift of the output signal is 90°. When the input capacitance of the amplifier, which is represented as the sum of three components,

$$C_{\text{tot}} = C_d + C_{\text{in}} + C_{\text{diff}},$$  \hfill (4)

differs from zero, where $C_d$ is the photodiode capacitance, $C_{\text{in}}$ is the input capacitance of the inverting input of the amplifier, and $C_{\text{diff}}$ is the differential capacitance between its inputs, the behavior of the FR and PFC changes significantly (Fig. 2b).

This is due to the fact that the feedback coefficient $\beta$ becomes frequency-dependent:

$$\beta = \frac{X_{C_{\text{tot}}}}{R_f + X_{C_{\text{tot}}}} = \frac{1}{1 + j2\pi f R_f C_{\text{tot}}},$$  \hfill (5)

as a result, the loop gain is described by the two-pole function:

$$A(f)\beta = \frac{A_0}{1 + \frac{f}{f_c} + \frac{f}{f_c'}}.$$  \hfill (6)

The appearance of an additional pole at the frequency $f_c = \frac{1}{2\pi R_f C_{\text{tot}}}$ leads to a decline of $A(f)\beta$ in the high-frequency region with a rate of 40 dB/decade and an additional phase incursion of 90° (Fig. 2b). The total phase incursion at the moment of intersection of the graphs $A(f)$ and $1/\beta$ becomes close to 180°. Because of this, the transient process acquires an oscillatory character and a resonant peak appears in the graph of the transfer coefficient of the $I$-to-$V$ transimpedance amplifier.

It is possible to prevent the transition of the amplifier to the oscillatory mode by reducing the total phase
incursion via reduction of the $R_f$ and $C_{\text{tot}}$ values. It is practically impossible to reduce the input capacitance to zero, whose main component is the capacitance of the photodiode. A decrease in the nominal value of the feedback resistance involves a drop of the transfer coefficient, which is undesirable.

The conventional method for stabilizing the operation of a TIA is based on the use of a correcting capacitance $C_f$, which is connected in parallel to the feedback resistor $R_f$ (Fig. 1). The correcting capacitance changes the behavior of the feedback coefficient $\beta$:

$$
\beta = \frac{X_{C_{\text{tot}}} + j2\pi f R_f C_f}{1 + \frac{1}{2\pi f R_f C_f}}.
$$

Figure 3 shows the graph of the amplifier PFC when using the correcting capacitance. Jointly with the feedback resistance, it forms an additional “zero” in the graph $\beta A(f)$ at the frequency $f_p = \frac{1}{2\pi R_f C_f}$; this zero reduces the resulting phase incursion.

Figure 4 shows the plots of the FP and PFC at a fixed $R_f$ and different $C_f$ ratings. In Fig. 4a, the gain $1/\beta$ reaches a plateau at a level of approximately $1 + \frac{C_{\text{tot}}}{C_f}$ much earlier than the point of intersection with the FR $A(f)$ of the amplifier, which corresponds to the condition $f_p \ll f_0$. As a consequence, the phase margin at their intersection point that corresponds to the frequency $f_0$ has a value that exceeds $45^\circ$. In this case, the graph $A(f)$ changes the decay rate from 40 to 20 dB/decade in the interval $f_0 - f_p$. A resonance peak in the transfer-coefficient curve $I_{\text{sig-to-V}}$ is absent or has a small value. When the gain $1/\beta$ reaches a plateau at the moment of intersection with the graph $A(f)$ at $f_p = f_0$ (Fig. 4b), the phase margin is $45^\circ$. A pronounced resonant peak that corresponds to a partial phase compensation appears on the $I_{\text{sig-to-V}}$ transfer-coefficient curve. At $f_p > f_0$ (Fig. 4c), there is almost no phase compensation, the phase margin is less than $45^\circ$, and the amplitude of the resonant peak actually coincides with its amplitude in the absence of a corrective capacitance.

In essence, $C_f$ stabilizes the operation of the TIA due to the narrowing of its operating frequency band. The optimal value of the correcting capacitor that cor-

\[ \beta \approx \frac{1}{1 + \frac{C_{\text{tot}}}{C_f}}. \]
The first component in expression (11) is shot noise, while the second is thermal noise. Both of these components can be described as “white” noise. In the case where the loop gain is \(|A(f)|β \gg 1\), the bandwidth of these components is limited by a low-pass filter, which is formed by the capacitance \(C_f\) and the feedback resistor \(R_f\), with the cutoff frequency \(f_p = \frac{1}{2\pi R_f C_f}\). If the input capacitance \(C_{\text{tot}}\) is large, the shot and thermal noises are then also frequency dependent [3].

The noise voltage reduced to the amplifier input is determined through the spectral density \(e_{\text{amp}}\) and the gain \(A_{\text{noise}}\):

\[
e_{\text{rmsamp}}^2 = \int_{0}^{f_{\text{GBP}}^2} A_{\text{noise}}^2(f) e_{\text{amp}}^2 df,
\]

where

\[
A_{\text{noise}} = \frac{1 + j2\pi R_f (C_{\text{tot}} + C_f)}{1 + j\frac{f}{f_0}},
\]

\(f_0 = \frac{f_{\text{GBP}} C_f}{C_f + C_{\text{tot}}},\) and \(f_{\text{GBP}}\) is the unity-gain frequency of the TIA.

In Fig. 6, a solid line marks the graph of the noise gain \(A_{\text{noise}}\), the dashed and dashed—dotted lines are the graphs of the transfer coefficient of the amplifier with the open feedback loop \(A(f)\) and the feedback coefficient of the amplifier \(1/β\), respectively. The noise gain is small and constant in region 1; therefore, we are mainly interested is the regions 2—4. In the frequency range from \(f_z = \frac{1}{2\pi R_f C_f + C_{\text{tot}}}\) to \(f_p = \frac{1}{2\pi R_f C_f}\), the noise gain increases with a rate of 20 dB/decade. In the region 4 where \(A_{\text{noise}} = A(f)\), the noise gain decreases with a rate of 20 dB/decade down to \(f_{\text{GBP}}\). In the region 3, it reaches a plateau and becomes equal to 1 + \(\frac{C_{\text{tot}}}{C_f}\). This allows us to consider the RMS noise voltage in the range from \(f_z\) to \(f_0\) or from \(f_p\) to \(f_{\text{GBP}}\) as

\[
e_{\text{rmsamp}}^2 \simeq \left(\frac{1 + C_{\text{tot}}}{C_f}\right)^2 e_{\text{amp}}^2 (f_z - f_0) \quad (14)
\]

When the TIA operates with large-area photodiodes, the predominant contribution to \(C_{\text{tot}}\) is made by the diode capacitance \(C_d\). As \(C_d\) increases, the noise gain increases as well. In the high-frequency region, beginning with

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*Fig. 5.* An equivalent circuit of the frequency-compensated TIA.

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 responds to a phase margin of 65° in the amplifier feedback loop is determined as

\[C_f = \frac{1}{2\pi (0.707) f_c R_f}.\]

Noises of the photodiode and TIA are factors that directly affect the resulting dynamic amplitude range of the detector. Due to the relatively small energy gap width of semiconductor photodiodes (units of volts) and the high energy of photons and other particles detected by them (tens of kiloelectronvolts—units/tens of megaelectronvolts), the number of charge carriers generated in the region of the \(p–n\) junction usually is large, thus determining a high value of the signal-to-dark current ratio. Therefore, the noise of the dark current of semiconductor photodiodes in most applications that are significant for plasma physics and CTF is not a factor limiting the operating dynamic range of the detector. The determining contribution to this limitation is made by amplifier noises.

Figure 5 shows an equivalent circuit of the frequency-compensated TIA with all noise sources. The frequency dependence of its output voltage is defined as

\[e_0(f) = i_{\text{sig}} K_s(f) + e_{\text{noise out}}\]

where \(i_{\text{sig}}\) is the signal current, \(K_s(f)\) is the transfer coefficient of the TIA, and \(e_{\text{noise out}}\) is the signal noise component at its output:

\[e_{\text{noise out}} = \sqrt{e_{\text{rms}}^2 + e_{\text{rmsR}}^2 + e_{\text{rmsamp}}^2} = \sqrt{R_f^2 q I_B f_p + 2\pi k T R_f f_p + e_{\text{rmsamp}}^2} \quad (10)\]

In this expression, \(e_{\text{rmsR}}\) is the root-mean square (RMS) value of the voltage noise component at the feedback resistor \(R_f\), \(e_{\text{rms}}\) is the noise determined by the noise component of the amplifier input current \(I_B\) that flows through the feedback resistor \(R_f\), and \(e_{\text{rmsamp}}\) is RMS value of the remaining noise components reduced to the TIA input.
noise that is caused by the input noise voltage $e_{\text{amp}}$ of the OA dominates in the TIA. Relationship (15) unambiguously indicates that for confident detection of broadband signals it is necessary to use photodiodes with the lowest possible capacitance $C_d$, as well as broadband OAs with $i_{\text{amp}} \ll \sqrt{2qI_{\text{sig}}} \approx \text{RMS}$, which have the noise voltage $e_{\text{amp}}$ and the input capacitance $C_{\text{tot}}$ with extremely low values reduced to the input. The resistor $R_f$ in the feedback circuit sets the value of the TIA transfer coefficient in the region of low and medium frequencies. It usually has a large nominal value and, therefore, has almost no significant effect on the amplifier noise characteristics. It is not always possible to fulfill the key condition, that is, to minimize the value of the parasitic capacitance that is connected to the amplifier input. The exceptions include detectors based on $p-i-n$ and avalanche photodiodes with a small active-zone area and broadband OAs.

CONSTRUCTION OF THE AMPLIFICATION PATH FOR DETECTORS WITH A SMALL JUNCTION CAPACITANCE

However, even for detectors with a small junction capacitance, it is extremely difficult to achieve the operating frequency band that is necessary for numerous applications, with an acceptable signal-to-noise ratio (SNR) at the amplifier output. This is explained by the characteristics of modern broadband OAs and the complexity of taking parasitic factors into account when designing TIAs based on them. In this regard, single-chip analogues of TIAs that have recently appeared and include OAs, feedback-loop elements, and correction circuits in their composition look more attractive. One example is the OPA857 amplifier (Texas Instruments), which, with a feedback resistor with a resistance of 20 kΩ and an input capacitance of 1.5 pF, provides the signal amplification in the operating frequency band from 0 to 105 MHz [4]. At the same time, the value of the noise current, which is reduced to the amplifier input, taking the component in the form $e_{\text{amp}}C_{\text{tot}}$ into account, does not exceed 15 nA (RMS). The OPA857 is intended to operate with avalanche or $p-i-n$ photodiodes that have a relatively small area of the active zone and a corresponding value of the junction capacitance $C_d$ of up to 4.7 pF. It is important that in the specified range of capacitance variation this amplifier is stable and its FR with the closed feedback loop remains almost flat. Only the value of the upper cutoff frequency of the bandwidth (from 105 to 80 MHz) and, in a relatively small range (from 15 to 23 nA), the value of the noise current reduced to the amplifier input change. The unique combination of these parameters makes it possible to use the OPA857 as part of broadband radiation detectors, e.g., those typical for diagnostics of Thomson scattering, which is intended to measure the density and temperature of the electron component of plasma in magnetic traps [5].

The key features of the Thomson-scattering diagnostics are as follows: the short duration of the probing pulse (1–20 ns) with the ratio of the probing and scattered radiation powers at a level of $10^{-15}$, the presence of powerful background plasma radiation, as well as laser radiation reflected from the walls of the vacuum chamber, which enter the detector aperture. In this case, it is possible to select a useful signal only when the measuring paths are operating in the oscillographic mode.

In particular, in the diagnostics of Thomson scattering on an axially symmetric magnetic gas–dynamic trap (GDT) of the Institute of Nuclear Physics (Siberian Branch, Russian Academy of Sciences) [6], a neodymium laser that generates a pulse with an energy of 1.7 J and a duration of ~10 ns is used as the radiation source. The laser operates in a pulse–periodic mode with a pulse repetition rate of up to 10 Hz. The optical diagnostic system is focused on measuring the plasma temperature at six spatial points. Each point has its own measuring path that includes a spectrometer with six photodetectors. The number of photons in a pulse of plasma-scattered probing radiation at each spatial point is approximately $(2-3) \times 10^4$, which corresponds to approximately $5 \times 10^3$ photons that enter the detector aperture. The value of the signal current of the detector, which uses an S11519–15 avalanche photodiode (Hamamatsu), is defined as

$$I_{\text{sig}} = \frac{M\eta e_{\text{pp}}}{t} \approx 2 \mu\text{A},$$  

where $M$ is the avalanche gain (25), $\eta$ is the quantum efficiency of the photodetector at the laser wavelength (1.06 μm), $e$ is the electron charge, $t$ is the laser-pulse duration, and $n_{\text{pp}}$ is the number of photons. Using these parameters, we can estimate the RMS value of the photodiode noise current:

\[ f_{\text{dom} \text{inate}} = \frac{1}{2\pi e_{\text{amp}}C_{\text{tot}}} \sqrt{2qI_{\text{sig}} + I_{\text{amp}}^2 + \frac{4kT}{R_f}}, \]  

\[ (15) \]
where $I_d$ is the dark current of the avalanche photodiode (it equals $\sim 9$ nA), and $f_{apd}$ is the cutoff frequency of its transfer coefficient ($\sim 100$ MHz).

Figure 7 shows a diagram of the amplification path of the Thomson-scattering diagnostic detector. It has three stages. The first amplification stage is based on an OPA857 TIA. The capacitance of the avalanche-photodiode junction that is connected to its input has a value of $\sim 3$ pF. Due to it, as well as the parasitic capacitances of the inverting input of the amplifier ($C_{tot}$), the noise current (RMS) reduced to this input does not exceed 18 nA in the entire operating frequency band (0–100 MHz). The value of the current noise component reduced to the OPA857 input, taking the noise component of the photodiode current into account, is equal to

$$I_{det} = \sqrt{I_{apd}^2 + I_{amp}^2} = \sqrt{63^2 + 18^2} = 65.5 \text{ nA}.$$  

(18)

It can be seen that the main contribution to $I_{det}$ is made by the noise component of the avalanche-photodiode current $I_{apd}$. It also determines the value of the resulting ratio of the signal amplitude to the RMS noise value, which characterizes the resolution of the detector unit:

$$S/N = \frac{I_{sig}}{I_{det}} = 30.5.$$  

(19)

This result is acceptable but not impressive. It should be noted that it corresponds to the operating frequency band of the detector (0–100 MHz) that is redundant from the physical point of view in the Thomson-scattering diagnostics. The fact is that in order to determine the plasma temperature in this diagnostics it is necessary to know the broadening of the spectral line of probing radiation, which is determined by its scattering by energetic electrons. There are two ways to solve this problem: by reconstructing the shape of the spectral line of scattered radiation from the amplitudes of the scattering signals, which are registered by detectors in a set of neighboring spectrometer windows, or from the integral values of these signals. Obviously, from the metrological viewpoint, the second method should be preferred, since within its framework the high-frequency components of noise and interference that are superimposed on the scattering signals are efficiently suppressed. From the radio-engineering point of view, the integration operation is equivalent to the procedure of filtering the high-frequency signal components; in our case, it is equivalent to reducing the upper cutoff frequency of the amplification-path bandwidth to an acceptable level. This operation is performed in all stages of the amplification path: in the TIA and the intermediate amplification stage, using corrective capacitances in the feedback circuits; in the output stage, due to its construction on the basis of an active two-pole filter. The cutoff frequencies of the FR of these links and elements are identical and equal to 60 MHz, thus providing the slope of the decline of the FR amplification path in the high-frequency region at a level of 80 dB/decade. In this case, the signal bandwidth is limited from above (at a level of $-3$ dB) to a frequency of 26 MHz, while the noise amplification frequency band is limited to a frequency of 29.5 MHz. These values are almost four times lower than the values of similar cutoff frequencies of the elements of the amplification path without correcting and filtering circuits. Due to this, the resulting SNR at the amplifier output is almost doubled: from 30 to 60. The subsequent procedures for processing detector signals are performed using their digitization and digital filtering paths. Figure 8 shows the output signals of the Thomson-scattering diagnostic detectors that were obtained in a real experiment with a probing-radiation pulse duration of $\sim 50$ ns. The time shift of signals relative to each other is due to the design of the spectrometer.
CONSTRUCTION OF THE AMPLIFICATION PATH FOR DETECTORS WITH A LARGE JUNCTION CAPACITANCE

When trying to use a photodiode with a larger junction area in the detector and, accordingly, with a larger parasitic capacitance $C_d$, the characteristics of the amplification paths constructed on the basis of TIAs deteriorate sharply. First, in proportion to the value of this capacitance, the amplitude of the output-signal noise component begins to increase; when this capacitance exceeds a certain critical value the amplifier loses stability and ceases to perform its functions. Therefore, it is possible to expand the bandwidth of the amplification path that operates in tandem with a photodiode with a relatively large junction capacitance only by changing the circuitry of this path, e.g., using a "cascode" connection of its design (Fig. 9), in which the functions of the photocurrent receiver are performed by a transistor connected according to a common-base circuit. The relatively low input impedance of this transistor together with the photodiode capacitance form a pole with the cutoff frequency $f_c = \frac{1}{2\pi C_d r_e}$ in the FR of the amplification path, where $r_e$ is the dynamic resistance of the emitter junction. The collector circuit of the transistor, whose load is the inverting input of the TIA, has a signal–current transfer coefficient that is close to unity and efficiently isolates this input from the photodiode capacitance.

In essence, in a cascode circuit, the photodiode capacitance at the input of a classical TIA is replaced by the collector–base junction capacitance of the transistor, which may have a value of fractions of picofarads. As a result, the TIA fully realizes the totality of its best characteristics: a wide bandwidth and a small value of the current noise component reduced to its input. The input transistor or, more precisely, the value of its quiescent current $I_0$, becomes the main element that determines not only the bandwidth but also the noise characteristics of the cascode amplifier. It determines the value of the input dynamic resistance of the emitter junction $r_e = \frac{\varphi}{I_0}$, where $\varphi = kT/e \approx 25$ mV at room temperature and the time constant of the input circuit $t = r_e C_d = \frac{\varphi}{I_0} C_d$. The collector noise current also depends on the quiescent current $I_0$ as $I_n = \sqrt{2eI_0 F}$. Here, $F$ is the upper cutoff frequency of the signal path bandwidth. The latter expression does not take the effect of the transistor-base resistance on the noise value into account, since its contribution at $I_0 \ll 1$ mA is negligibly small. The values of the transistor noise current $I_n$ and its quiescent current $I_0$ are interrelated via the simple dependence:

$$I_n = I_0 \sqrt{\frac{e}{\pi \varphi C_d}} = 1.42 I_0 \times 0.001 \frac{1}{\sqrt{C_d}}, \quad (20)$$

where $C_d$ is the value of the photodiode capacitance in picofarads. It follows from this ratio that with a photodiode capacitance of 10 pF and an RMS value of the transistor noise current comparable with the TIA noise current (15 nA) the $I_0$ value must not exceed 36 μA. In this case, due to the relatively large value of the transistor input resistance ($r_e \approx 700$ Ω), the upper cutoff frequency of the input-path bandwidth is limited to 23 MHz. At the cost of increasing the quiescent current of the transistor to 100 μA, it can be increased to 64 MHz with a simultaneous increase in both the noise current component of the input stage (up to 45 nA) and the resulting value of the noise current (RMS) (reduced to the amplifier input) to 47 nA. With an amplitude of the photodiode signal current of 10 μA, the resulting SNR reaches approximately 212, which is already acceptable for many applications.
Thus, the cascode amplification circuit is efficient when working with photodiodes with a capacitance of ~10 pF. The possibility of using it with photodiodes with a larger junction capacitance is still questionable. Let us try to modify this scheme so as to provide the solution of the formulated task. It is obvious that in order to reduce the influence of the photodiode capacitance on the characteristics of the amplifier, it is necessary to get rid of this very capacitance at least partially if not completely. This can be done by prohibiting its recharging with a signal current. To do this, it is sufficient to stabilize the voltage at the photodiode with the help of auxiliary elements, e.g., with the help of a voltage follower and a signal-level offset circuit that maintain the equality of the voltage drops at both its terminals (Fig. 10).

Such a solution was considered in [7]. It is based on the transfer of the voltage change from the emitter of the input transistor, to which the signal output of the photodiode is connected, to its second output. For this purpose, an emitter voltage follower based on the $Q_2$ transistor and an $R_3$, $R_4$, $C_2$ resistive-capacitive circuit for biasing the level of its output signal are used. These elements prevent a charge exchange of the parasitic capacitance $C_d$ of the photodiode in the frequency range in which the value of the $Q_2$ transistor output resistance is much lower than the reactance of this capacitance: $r_e \ll X_{C_d}$. Therefore, with a fixed value of the photodiode capacitance, it is possible to expand the operating frequency band of the modified cascode amplification circuit only by reducing the value of $r_e$, which is equivalent to an increase in the quiescent current $I_{Q2}$ of the $Q_2$ transistor. It should be taken into account that an increase in the quiescent current $I_{Q1}$ inevitably leads to an increase in the base current $I_{Q2}$ of the $Q_3$ transistor, which, because of the necessity of stabilizing the value of the input-stage noise component at a level of $\sqrt{2eI_{Q0}F}$, must be much smaller than the quiescent current $I_{Q0}$ of the input stage. This contradiction is resolved when constructing a follower based on a bipolar microwave transistor with a high current gain or on a low-noise high-frequency field-effect transistor (FET) with a high slope, e.g., ATF55143. In the case of a bipolar transistor, the operating frequency band of an amplifier based on the modified cascode circuit increases by a factor of $I_{Q0}/I_{Q1}$ in comparison to its previous version with preservation of the spectral density of the noise component at a level close to $\sqrt{2eI_{Q0}}$ up to the frequency

$$F = \frac{\sqrt{I_{Q0}}}{2\pi r_{Q1} C_d}.$$  

Starting from this frequency, the quiescent current of the $Q_2$ transistor will begins to make a decisive contribution to the current noise component reduced to the TIA input. Similar relationships, when replacing $r_e$ with $1/s$, where $s$ is the slope of the transfer characteristic, are also valid in the case of a signal follower based on a FET.

A modified cascode photocurrent amplification circuit was used at the GDT installation in the diagnostics of the intensity of the proton flux generated in a nuclear reaction: $D + D = p + T + 4.032$ MeV.

The corpuscular diagnostics of plasma is based on determining the flow intensity of particles that are produced during the interaction of plasma with beams of neutral atoms injected into it. The specific feature of the GDT installation is that the concentration of trapped particles in the central probkton is relatively small; however, their density increases significantly in the region of the magnetic mirrors. As a consequence, the intensity of particles that are detected by the diagnostics may vary in a wide range, which implies the necessity of operation of the measuring paths both in the counting mode (at a low plasma density) and in the oscillographic mode (at a high plasma density), in which the behavior of the flow intensity of particles that leave plasma is recorded in time.

In the corpuscular diagnostics in the GDT installation, protons are detected by a D1A silicon diode manufactured by SNIIP-PLYuS (Moscow, Russia) with a surface area of 100 mm$^2$, which has a capacitance of the reversely biased junction of ~100 pF at a bias voltage of 40 V. Due to the intense interaction of protons with the residual gas, which forms a “coat” of the plasma filament, and with the material of the walls of the vacuum chamber, the diode is located together with the collimator in the vacuum volume. The diode
is connected to the amplifier—shaper (Fig. 11), which is placed outside of this volume in a shielded box, using a vacuum-tight connector and connecting wires with a length of up to several centimeters.

At an energy of formation of an electron-hole pair of 3.66 eV, each proton that is formed in plasma within the framework of the $D-D$ reaction generates a charge of approximately 100 fC in a silicon diode. Such a small value of this charge, a relatively large capacitance of the diode, and its severe operating conditions are the key factors that determine the amplifier—shaper circuitry.

This amplifier consists of two series-connected components: a modified cascode amplifier (Fig. 12) and an output amplifier—shaper $A_{\text{out}}$ based on an OA with differential inputs and outputs. The upper cutoff frequency of the LMH3401 with a parasitic input capacitance at a level of ~1.5 pF and an equivalent feedback resistance of 20 kΩ reaches 250 MHz. In this frequency band, the noise current (RMS) reduced to its input does not exceed 49 nA. The input stage of the cascode amplifier is based on a BUF740 low-noise microwave transistor ($Q_1$) with the quiescent current $I_{0Q1} = 25 \mu A$; the stage for stabilizing the photodiode voltage is based on a high-frequency BC847 transistor ($Q_2$) with an emitter current of 1 mA.

The operating frequency band of the cascode amplifier, in which the spectral density of the current noise component reduced to the input is constant and is determined by the value of the current $I_0$, is limited from above by the value

$$ F = \frac{I_c}{2\pi r C_d} = \frac{\sqrt{1}}{2 \times 3.14 \times 25 \times 10^{-9}} = \frac{6.32}{6.28 \times 25 \times 10^{-9}} = 40 \text{ MHz}. $$

In this frequency band, the value of the input—current noise component for the entire amplifier, taking the above estimates for each stage into account, does not exceed 27 nA (RMS), which corresponds to a SNR at a level of 370. The problem is that the operating frequency band of the stage with a common base and the TIA is much wider. Therefore, in order to obtain the desired result and exclude the influence of the quiescent current of the emitter follower and the high-frequency components of the input-transistor and TIA noise current on the resulting SNR, the operating frequency band of the amplification path is limited using an auxiliary filter. The role of the filter is performed by an output shaper that is based on a differential amplifier. Like an active filter of the detector

![Fig. 11. The detector of the proton-flux intensity.](image1)

![Fig. 12. The circuit of the amplifier—shaper for the detector of the proton-flux intensity.](image2)
of the Thomson-scattering diagnostics, it restricts the bandwidth of the operating frequencies of the amplification path from above to the cutoff frequency $F$ and increases the transfer coefficient of the latter by 5 times. The specified cutoff frequency is set by the resistive–capacitive elements of the feedback circuits of the amplifier–shaper output stage.

Figure 13 shows a signal response of the described amplifier to a particle with an energy of ~5 MeV that enters the detector aperture. The figure also shows an oscillogram of the signal response to the arrival of the same $\alpha$ particle at a detector with a similar amplifier–shaper and a D4.5A silicon diode, which has a fourfold active-region area and, thus, the corresponding value of the parasitic capacitance $C_d$ of the junction.

CONSTRUCTION OF AN AMPLIFYING PATH FOR DETECTORS LOCATED REMOTELY

Unlike silicon detectors, artificial-diamond-based detectors are radiation-resistant and operate at high bias voltages (hundreds of volts); due to the high mobility of carriers and the small width of the $p–n$ junction, they have subnanosecond durations of signal responses and an energy resolution acceptable for thermonuclear applications (at a level of percent or less). Carbon atoms of the diamond detector have a relatively large cross section of the interaction with high-energy neutrons. Due to their design features, detectors based on them are characterized by a much smaller interaction cross section with energetic $\gamma$ quanta. Because of this, diamond detectors are used to register neutron events on the installations characterized by severe radiation conditions and, accordingly, intense thermonuclear neutron fluxes. WEST (France, Cadarache), JET (UK, Culham), and ITER (France, Cadarache) are examples of such installations. Together with collimators and electromagnetic protection elements, the detectors are placed in close proximity to the experimental complex and are interfaced with preamplifiers–shapers, which are placed behind the biological-shield elements, and lines based on a radiation-resistant cable with a length of up to several tens of meters (Fig. 14).

The bias voltage is fed along the same lines to the detectors from the corresponding power sources through ballast resistors. The inputs of the preamplifiers are isolated from the cable lines and DC bias-voltage sources via decoupling capacitances [8].

The total charge that is generated by a thermonuclear neutron (14 MeV) as a result of nuclear reactions in a diamond detector is relatively small, slightly larger than 100 fC. When passing through a long cable communication line, the initial subnanosecond signal response of the detector is transformed due to dispersion into a current pulse with a duration of 10–15 ns, which determines the possibility of its operation with neutron fluxes of rather high intensity ($\geq 10^7$ events/s). This option is realized if the integration time constant $\tau$ of the amplifier–shaper is comparable with the characteristic current-pulse duration at the output of the cable line and its input resistance $r$ is equal to the characteristic impedance of this line. These conditions determine not only the high cutoff frequency of the preamplifier FR but also the need to eliminate reflections in the signal path and, as a consequence, a relatively large value of the spectral density of the noise current generated by the matching resistor or its equivalent at the receiving end of the cable communication line. When the line is matched using a 50-Ω resistor, the spectral density of the noise current has a large value, $i = \frac{4kT}{R} = 18$ pA/$\sqrt{\text{Hz}}$. On an integration
interval whose duration is comparable with the typical pulse duration at the cable-line output (15 ns), it forms a noise charge of ~2.6 fC, which is only 35–40 times smaller than the charge formed by a thermonuclear neutron when it hits the diamond detector. As a consequence, the energy resolution of the signal path turns out to be poor. The situation can be somewhat improved by replacing the resistive matching line of the cable with its less noisy counterpart. The functions of the latter can be performed by the dynamic resistance $r_e$ of the base–emitter junction of the transistor connected according to the common-base circuit (Fig. 15).

Due to the dependence of the dynamic resistance of the transistor emitter–base junction $r_e$ on the emitter current $I_e$ ($r_e = \varphi/I_e$, where $\varphi$ is the temperature potential), this matching at the transistor emitter current $I_e = 0.5$ mA allows the noise determined by resistive matching to be reduced by a factor of $\sqrt{2}$. It can be seen that the cascode circuit of the signal amplifier—former in tandem with the diamond detector has certain advantages, as in previous cases. It reduces the level of the noise component reduced to the input of the amplifier—shaper, but does not ensure the input-resistance stability. This is due to the fact that the dynamic resistance of the emitter–base junction of the transistor is modulated by the signal current of the diamond detector. As a consequence, the level of its reflection at the receiving end of the cable line changes as well, depending on the signal amplitude. In view of the random nature of signals that are generated by the diamond detector, it can be concluded that the level of reflections directly affects the resulting SNR, thus leveling the gain obtained via changes in the cable-line matching scheme. Thus, the task of minimizing the modulation level of its input resistance by the signal current becomes urgent in the signal amplifier—shaper of the diamond detector. The input stage based on transistors of different conductivities $Q_1$ and $Q_2$ that are connected according to the circuit with a common base allows this problem to be solved (Fig. 16).

A signal from the cable line is fed to the emitter–base junctions of these transistors through decoupling capacitances that isolate them from the bias-voltage source. These junctions are connected in parallel with respect to the signal source. The dynamic resistance of each junction with a value of 100 $\Omega$ is observed at a quiescent current of the emitter of $I_e = 0.25$ mA. Changes in the signal current in the emitter circuit of the transistors are the same and equal to $I_e/2$ but are opposite in sign. Therefore, the resistances of the emitter–base junctions of the transistors change under the influence of a signal current by almost the same value but in the opposite directions. In the transistor in which the signal current is added to the quiescent current this resistance decreases; in the second one, where the signal current is subtracted from the quiescent current, this resistance increases. Since the transistor base–emitter junctions are connected in parallel with respect to each other, these changes in their dynamic resistances actually compensate for each other. As a consequence, the modulation of the amplification-path input resistance by the signal current is efficiently suppressed. The value of the SNR remains unchanged. This is due to the fact that half the quiescent currents in a two-transistor matching circuit together form the same noise current as the quiescent current of its single-transistor analogue. A relatively low SNR at the input and large fluctuations in the position of the zero line of signals at the amplification-path output, which depend on the detector loading, are the key factors that affect the resulting energy resolution of the neutron diagnostics. It is possible to reduce the influence of the first factor, as noted earlier, by filtering the high-frequency components of the amplification-path output signal and the second one by eliminating the “averaged” component from this signal on the interval of its registration. The latter operation is equivalent to eliminating the low-frequency spectral components of the signal from it. An obvious conclusion suggests itself: it is desirable to
perform the basic procedures for processing the output signal of the amplification path of neutron diagnostics with its digital equivalent, which is represented in a spectral or amplitude–frequency form, which is essentially the same.

Let us return to the amplification path. In order for the above processing procedures to become real, the path must generate an output signal with the maximum possible SNR, which is suitable for digitization by a high-speed ADC. Since the result of the neutron diagnostics is represented in the form of an energy spectrum of particles or the flux intensity of neutrons with a certain energy that fall within the detector aperture, two key factors have a decisive influence on the resulting characteristics of the amplification path. These are the maximum permissible load of the detector and the maximum possible duration of its output signal, which excludes a sharp increase in the dead diagnostic time due to overlapping events. When the detector load is several million events per second, which is typical of the neutron diagnostics of many plasma facilities, the characteristic duration of an output signal of the detector amplification path lies in the range of 20–70 ns (FWHM). Wider signals are preferred because the bandwidth limitation of the amplification path that is necessary for their formation in the high-frequency region is transformed into an increase in the resulting SNR. In addition, in the case of “long” signals, the requirements for the performance of ADCs that convert the current amplitude values of these signals into a digital equivalent are considerably attenuated.

The resulting duration of the output signal and the SNR depend on the ratings of the corrective capacitances. Figure 18 shows a characteristic waveform of the output signal of the amplification path of the neutron diagnostics.

CONCLUSIONS

The task of constructing the amplification paths for semiconductor detectors of plasma diagnostics is not trivial despite its apparent simplicity. The final signal/noise characteristic of the amplification path and its frequency characteristics are significantly influenced by the parameters of the TIA (the input current, input voltage, bandwidth, etc.), as well as the input capacitance of the detector, the differential capacitance, and the input capacitance at the noninverting amplifier input. The methods for solving this problem strongly depend on the goals that the developer seeks to achieve: an increase in the bandwidth or an increase in the SNR. In most cases, it is necessary to come to a compromise between the allowable bandwidth in a particular task and the corresponding SNR value.
BROADBAND SIGNAL AMPLIFICATION PATHS FOR SEMICONDUCTOR RADIATION

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CONFLICT OF INTERESTS

The authors declare that they have no conflict of interests.

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