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VIRGO

Four quadrant diode front end module for the Virgo Linear Alignment 3/ 30 mW, *plus* configuration

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Preface:

Due to an upgrade, in Virgo, new four quadrant diode front ends for use of the Virgo Linear Alignment must be produced. In these new modules the orientation of the four quadrant diode is changed to the so-called *plus* configuration. Two types are designed: a 3 mW and a 30 mW type. The power handling capability is defined such that the total light power falling equally on two of the quadrants just saturates the amplifier. Extra care is taken for the cooling of the four quadrant detector.
The new modules are mechanically compatible with the existing modules. Compared with the spare modules they are 3 cm longer; 20 cm without connectors. A simple overload indication in the form of LED's is added. The HF overload indication is now connected to the amplifiers, instead of looking at the subtracted signal. An extra output is added to monitor the total HF signal (sum of quadrants).
This description may refer to other files, these can be found starting at:
http://www.nikhef.nl/pub/departments/et/virgo/

Mechanics:

The outer dimensions of the metal box are 48 * 100 * 200 mm. The mounting tube of 30 mm diameter is fixed 5.5 mm below the centre of the front. This tube is 30 mm long. The four quadrant diode is mounted in the centre of the tube. The tube is insulated to prevent ground loops.

Connectors:

All connectors are located on the back of the module, opposite to the four quadrant diode.

- Power: 12 V : LEMO EPL.1S.302.HLN, Cable part: FFA.1S.302.CLAC27
  1: +12 V, 1A
  2: Gnd.
- HF outputs, horizontal, vertical and total: SMA-50
- DC outputs, four times: LEMO ECP.2S.308.CLV, Cable part: FFA.2S.308.CLAC62

+ configuration						
From front	**Old names**			**PCB name**	**8p LEMO**	
	Virgo	**INFN**	**4Q**		**Signal**	**Gnd**
Top right	DC4	D	R	DQ1	8	7
Top left	DC1	A	T	DQ2	6	5
Bottom left	DC2	B	L	DQ3	2	1
Bottom right	DC3	C	B	DQ4	4	3

- Bias control/ monitor: LEMO EPL.1S.303.HLN, Cable part: FFA.1S.303.CLAC27
  1. Monitor: -0.04 * V_{bias}.
  2. Control, 0…5 V.
  3. Gnd.
- Temperature monitor: Two pin connector, 0.1". Left is ground, right (near "Tmon") gives 1 mV/ K.

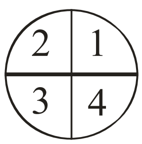
Electronics:

Four quadrant diode

The four quadrant diode is a YAG 444-4AH, an 11.3 mm diameter SI PIN diode. This series is IR enhanced. The active area is circular and the gap between the active areas is very small (app. 0.01 mm). The efficiency at 1064 nm is app. 0.45 A/W. The breakdown voltage is guaranteed at 200 V, so it can safely run at a bias voltage of 130 V. This bias voltage mainly reduces the

capacitance, thus improving the noise behaviour at higher frequencies. It marginally contributes to the collection efficiency.
The YAG 444-4 was produced by EG&G, which no longer exists. The device can now be obtained through Perkin Elmer. PE specifies a spectral noise current density of $120 \cdot 10^{-12}\ {}^{A}\!/\!_{\sqrt{Hz}}$, where EG&G specifies $2 \cdot 10^{-13}\ {}^{A}\!/\!_{\sqrt{Hz}}$.
Shot noise from the photodiode is caused by the stochastic collection of electrons and holes from leakage I_D (dark) and photo current I_P.
Using $I_{N,RMS} = \sqrt{2q \cdot (I_P + I_D) \cdot \Delta f}$ or $\overline{I_N} = \sqrt{2q \cdot (I_P + I_D)}$, with a dark current of 75 nA, we get $\sqrt{2 \cdot 1.6 \cdot 10^{-19} \cdot 75 \cdot 10^{-9}} = 1.2 \cdot 10^{-13}\ {}^{A}\!/\!_{\sqrt{Hz}}$. This agrees with the EG&G figure.
Centronics produces a similar device called QD100-4X, with the same diameter. Other types measure 8 to 20 mm.

Orientation



looking into the front of the module

Bias voltage

Reverse biasing the detector has a small positive effect in the charge collection of the electron-hole pairs, generated by the incident light. Linearity over many –seven or more- decades is obtained. However, this bias strongly decreases the detector's capacitance thus improving the behavior of the detector-amplifier combination. This in terms of gain-bandwidth, as well as in noise figures. The capacitance near zero bias is hard to define. But, comparing the values, going from a few Volts reverse bias to more than 100 V, the capacitance decreases by a factor of nearly 10. Increasing the bias voltage further has no use, since a plateau is reached at full depletion (100 to 150 V). The absolute limit is the break-down voltage, which is about 200 V. Also the dark current increases with the bias voltage. So, when the capacitance no longer noticeably decreases, one should not increase the bias voltage any further.

The bias voltage is internally generated from the 12 V supply. It is externally controlled by a control voltage, coming from a DAC. The input range is 0…+5 V. This is a linear control. 0 V gives the minimum output (<1 V) and 5 V gives full scale. Full scale is 130 V. When the bias cable is disconnected, the bias voltage is switched on. One can use a connector with a connection between pins 2 and 3 to disable the bias voltage.
The bias voltage is monitored at the same Lemo connector that is used for the control. The amplification is -0.04 times, thus giving a 0 to +5 V output. The bias voltage is generated by a high voltage module from Ultravolt. This module can deliver an output current up to 30 mA. For the 3 mW version this is limited to 2 mA by extra circuitry. The 30 mW version is limited to 20 mA. The actual current can be calculated from the sum of the DC outputs. The only way to see whether the current limit is active, is to monitor the bias voltage.

Power

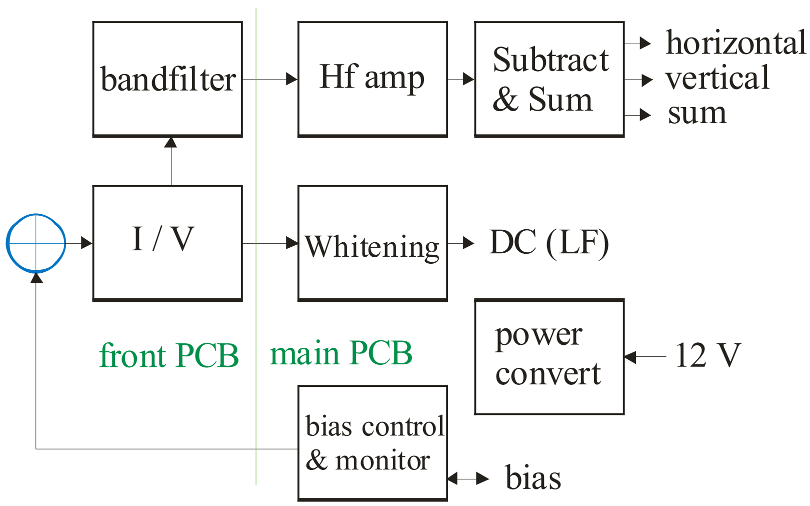
The single 12 V power supply should be capable of delivering 1 A. Without incident light, the module draws app. 300 mA. This goes up to 500 mA at full load (30 mW incident light).

Temperature monitor

The temperature of cooling block directly behind (thermally coupled to) the four quadrant sensor is monitored by means of an AD592ANZ, an integrated circuit temperature transducer. This gives 1 mV/K at the output pins. So take the mV reading and subtract 273. There is [will be] a separate document on the thermal behaviour of the module with respect to the incident light power.

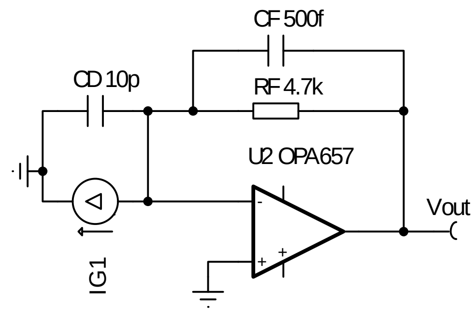
Amplifiers

The total amplification can be split into two stages: the transconductance amplifier, connected to the four quadrant diode and the various filters and output drivers. The over-all amplification of the DC outputs is set to generate 6 V when half of the expected incident light (3 mW or 30mW) falls on a single quadrant. The HF output level is approximately the same as the original Frascati modules. In the plus configuration more quadrants contribute to the signal however. Therefore the amplification of the HF amplifiers is slightly reduced compared to the *spare* modules.
The green line indicates the split between the detector board and main board.



Transconductance amplifier

Compared to the original Frascati design, some amplification has moved to the first stage. This improves the noise behaviour. The amplification linearly increases with the feedback resistor value, while the noise increases with the square root of this change. There are two limitations here. The first one is the gain bandwidth product of the amplifier (OpAmp). The second one comes from the feedback capacitor that is needed to suppress the zero that arises from the diode's capacitance in combination with the phase shift of the amplifier at high frequencies. So the feedback capacitor is not intended to limit the bandwidth, but is needed for stability. From the definition that 3 mW on two quadrants may saturate the system we come to the following: 1.5 mW * 0.45 A/W = 0.675 mA. The first stage of OpAmps saturates at 3.2 V. So the transconductance resistor becomes 4.7 kΩ. This, to get the maximum output at the rated optical power. For the 30 mW version this is 470 Ω.

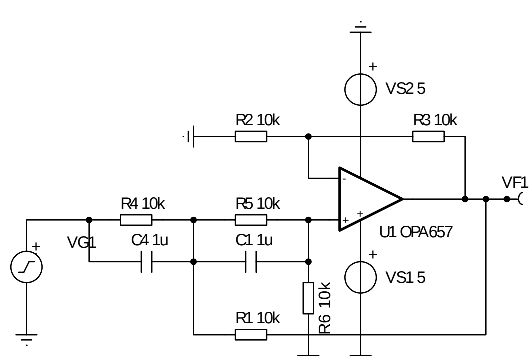


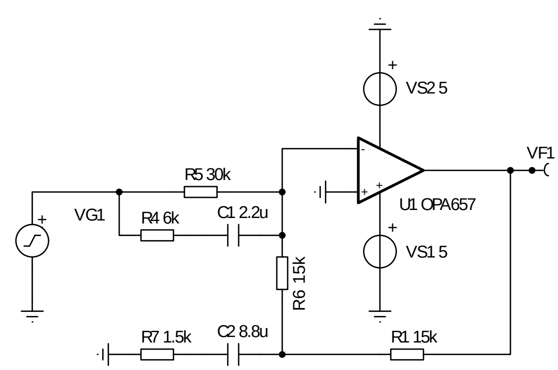
The feedback resistor, determining the conversion factor, is R_F. C_D is the photo diode's capacitance. C_F is the feedback or compensation capacitor, which prevents gain peaking due to the detector's capacitance. For a stable (flat gain) configuration a maximally flat 2nd order Butterworth frequency response is chosen. For this a feedback pole is needed at $\frac{1}{2\pi R_F C_F} = \sqrt{\frac{GBP}{4\pi R_F C_D}}$ or $C_F = \sqrt{\frac{C_D}{\pi GBP \cdot R_F}}$. Where GBP is the gain-bandwidth product of the amplifier. So, choosing R_F determines C_F. This sets the system bandwidth a little lower than needed for stability, but gives a very predictable behavior. With C_D = 10 pF, GBP = $1.6 \cdot 10^9$ (OPA657) and R_F = 4.7 kΩ, C_F becomes 0.65 pF. This is nearly covered by the parasitic capacitances already, but in practice an additional capacitor was needed. A capacitor with the value 0.68 pF is installed here. For the 30 mW version (with a 470Ω feedback resistor) this is 4.7 pF. The OPA657 showed some instabilities in this case and was replaced by an ADA4899-1.

The theoretical bandwidth of the whole circuit is $f_{-3dB} = \sqrt{\frac{GBP}{\pi R_F C_D}} \approx 100 MHz$ This clarifies the need of a much larger GBP than one would expect at first glance. With the feedback capacitor of 0.68 pF plus the parasitic capacitance the bandwidth is reduced to ~50 MHz.

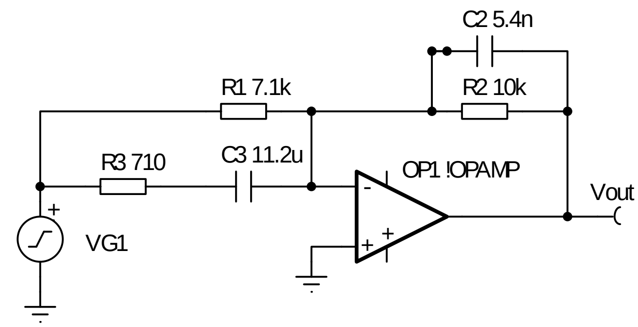
DC outputs

To leave some room for the low frequent AC signals we amplify the signal from the transconductance amplifier (limited to 3.2 V) by a factor of two, giving a full scale of 6.4 V for the DC component. In the *design change request* a whitening filter is specified. The amplification must increase by 40 dB above 10 Hz (double pole) and drop off above 2 kHz again. Second order whitening filters can be created in a few ways. Two passive sections are used around an amplifier, which provides positive feedback. The amount of feedback determines the Q of the filter. Below two examples of such filters are shown.

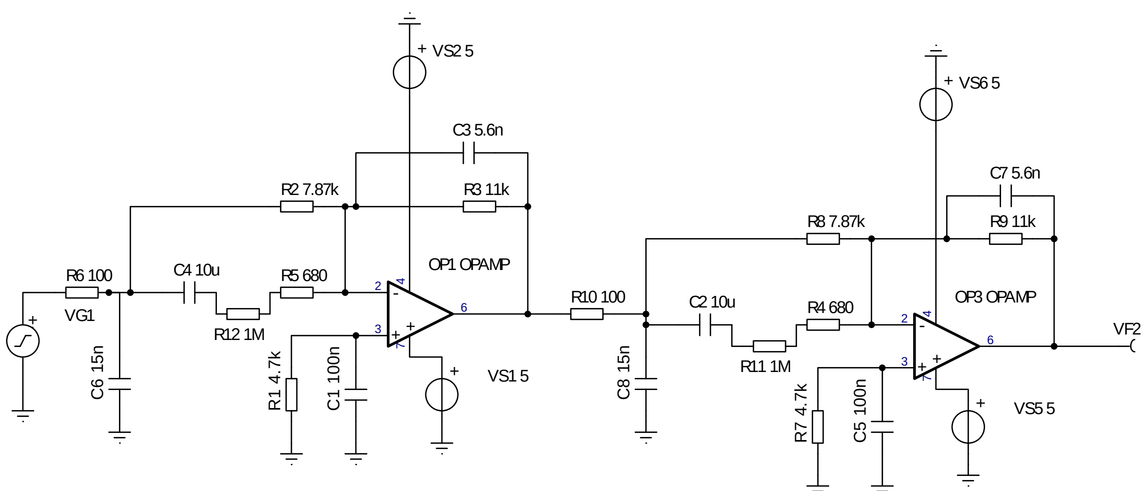




These types of circuits have a rather complex description and changing one specification has an impact on many components. Still a low-pass filter is needed, since it cannot be integrated in the same circuit. A Q of 0.5 however, as specified in the *design change request*, is a special case. This is equal to two stages of passive filters (simple poles or zeroes) decoupled by amplifiers to prevent the sections loading each other. By using this structure it is much easier to define the DC amplification, the AC amplification and the poles and zeroes separately. Below a single stage is shown. Choosing R2 determines all other components.



Here R_2/R_1 determine the DC amplification. R3·C3 determine the pole, with R_2/R_3 the amplification beyond that point. Amplification drops again at $R_2 \cdot C_2$, app. 2 kHz. Then we adjust this to practical values. Passive low pass stages are added to prevent feed through along the passive components where the OpAmps loose gain at higher frequencies. The 1 MΩ resistors represent the jumpers implemented in the actual design.

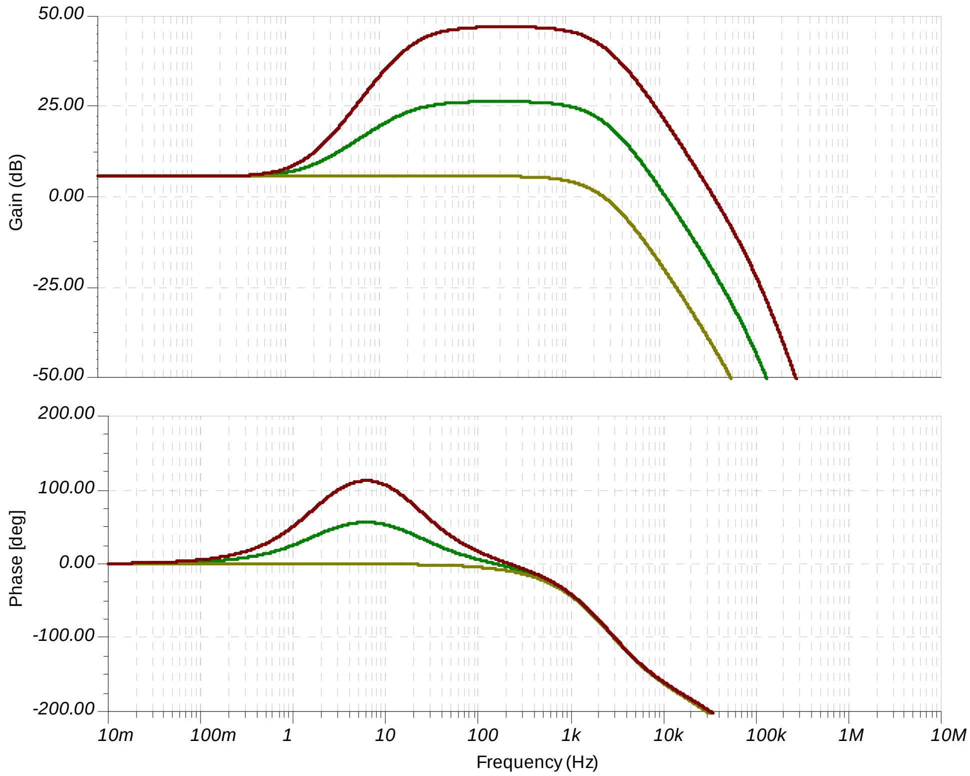


The jumpers (R11, R12) enable the option to three characteristics, as shown below. Resistors can be specified at 1 % tolerance, but capacitors at these large values have a larger tolerance. The capacitors are measured and sorted out. They are specified for each module in the file *module types*.

With the simple calculations from above, the component values are determined and corrected to values that are available. Then the circuit is simulated with results shown. Actually measuring the curve in a working module is not very accurate. The roll-off at 1.75 kHz goes rather well. Using a digital oscilloscope one measures the amplitude of the signal at 200 Hz. Then find the -3 dB point and read the frequency again. This should be done a number of times to get an impression about the accuracy. In between one does other measurements. The simulator result is 1.69 kHz. Where the filter rolls off at the low end is harder to determine. For the 40 dB version for 8 measurement an average of 33.7 Hz was found with a standard deviation[1] of 1.6 (extremes at 32.2 and 36.7). Simulation gives 31.6 Hz. The value of the capacitor in the simulation was 9.7 µF. This the average

[1] Not enough measurements for a correct standard deviation

of the measured values. Measuring the +3 dB point with respect to the DC amplification is even more difficult. DC here is in practice 0.1 Hz. The problem here is that the synthesized sine has frequency components where the filter already has a strong gain. This gives spikes on the output. Here an average of 1.36 was found with a standard deviation of 0.02 (extremes at 1.33 and 1.39). The simulator result is 1.25 Hz.



The references of the four outputs is the signal ground. On board there is an option to put 50 Ω resistors in these lines. This would create semi-differential outputs. So, it is not differentially driven. But the impedances seen by externally introduced signals are equal and the noise introduced can thus effectively be reduced by using a differential receiver.

HF outputs

The HF amplifiers are coupled to the transconductance amplifiers via band filters. These 6.26 MHz filters remove unwanted frequency components such as the double frequency, which is nearly as large as the base frequency in some locations. Though it is not likely that these harmonics saturate the HF amplifiers, it is good practice to remove frequency components you do not need for your measurement. The band filters are four channels of passive sections on the front or detector board. Though the Q is not very high, app. 8, these sections have to be trimmed to obtain a reasonable equal phase rotation.

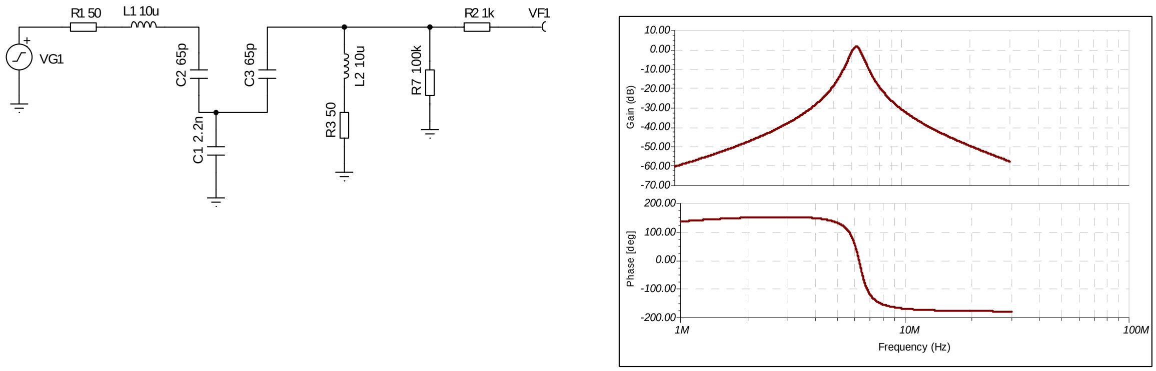


Figure 1: band pass filter

Approximate values, with C = 65 pF and L = 10 μH:

$$f_0 = \frac{1}{2 \cdot \pi \cdot \sqrt{LC}} = 6.26MHz$$

$$Q = \frac{1}{R} \cdot \sqrt{\frac{L}{C}} = 7.8 \Rightarrow B = 800kHz$$

$$Z \approx \frac{L}{C \cdot R} = 3k\Omega$$

Testing the board, it turns out that the Q of the coil is about 20. This reduces the pass band attenuation to ~-3dB. In simulation this is accomplished by adding resistors in parallel to the coils with a value of 8 kΩ. Also the trace capacitance is approximated (the 10 pF). This is shown below. The capacitors therefore change slightly.

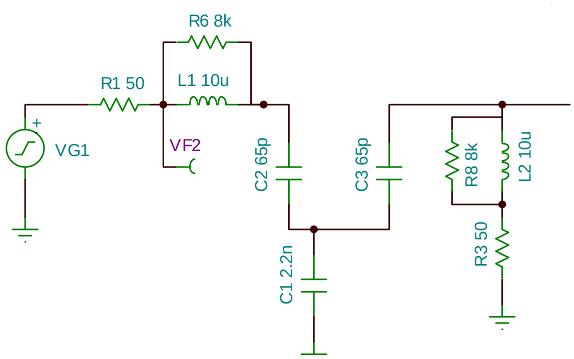
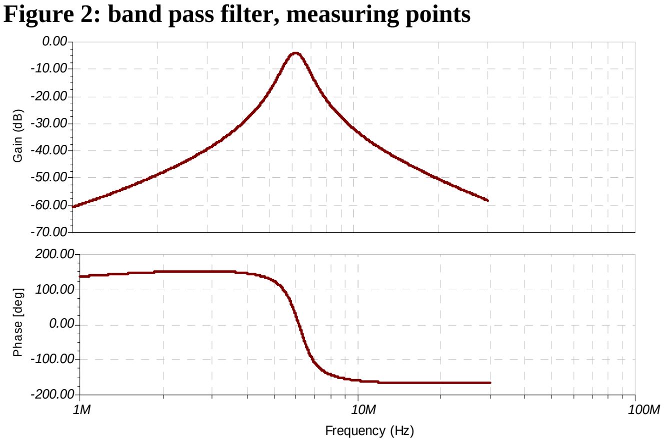


Figure 2: band pass filter, measuring points



The 63 pF capacitors are in fact a fixed 47 pF capacitor and a trimable capacitor of 4 to 25 pF. To observe the influence of these trimmers, an extra measuring point in the simulation is introduced, right after the first resistor in the chain. This shows that C2 is dominant in coupling energy into the filter and C3 optimizes in coupling to the next stage. The latter hardly changes the peaking frequency. So adjust C2 for maximum output, while injecting the desired peaking frequency, then adjust C3 for maximum output. In three or four rounds this should give an optimum. The phase shift is not really zero due to the loss in the inductors. For the 8 MHz types, the coils are changed to 5.6 µH. The base coupling is increased by lowering the capacitor value to 1.5 nF. This, to obtain the same loss as in the 6 MHz version.
The measured bandwidth is 850 kHz, which is close to the simulation value.

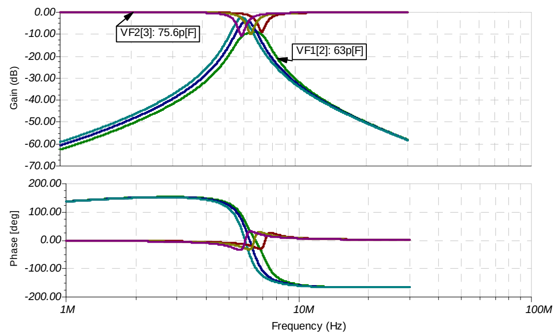


Figure 3: adjusting C2

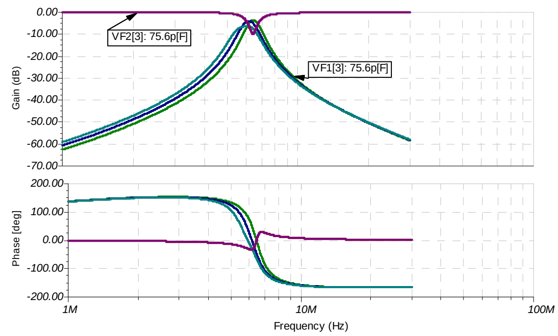


Figure 4: adjusting C3

All four channels are trimmed this way. Since the components of the four channels are not exactly the same, one has to find an optimum. After trimming individually, the output of one of the HF amplifiers is taken as a reference and the phases of the other channels are adjusted to be the same. The phase is more important than the amplitude. This, because the signals are subtracted after amplification.

If the band filter not required, it can be removed. On the detector board an AC coupling to the HF amplifiers is made. The input resistor to the filter and the 0 Ω output resistor are removed. The transconductance stage is coupled to the HF amplifiers by a 3.3 nF capacitor. On the main board the input resistor of 100 kΩ is replaced by 1 kΩ. This biases the OpAmp instead of the inductor on the front end board. This gives a pole at ~40 kHz. The RF outputs drop off again above 20 MHz. This is due to the transformer, not the amplifiers.

Compared to the original design, eight times magnification would be needed to get the same signal. However, in the + configuration, the signals from all quadrants are combined for both outputs against two in the **x** configuration. The amplification of this stage is therefore set to app. six times.

The vertical- and horizontal signals are determined by adding and subtracting the signals of the four quadrants. For the vertical signal this is the two top quadrants minus the two bottom quadrants. For the horizontal signal this is left – right (as seen from the front of the module). So the signal increases when the light spot moves to the left when looking towards the front of the module. This is done to be compatible with the existing definition.
The subtraction sections are followed by transformers thus isolating the cable shield from the electronics ground. The outgoing shield is capacitively coupled with the housing for EMI reasons. The output impedance is 50Ω.
A modified detector board is used to check the adder/ subtractor part. A 6.26 MHz signal is injected with an amplitude that gives 2 V RMS at the output of the HF amplifiers. The resulting signal is measured at the SMA output. The results are listed in the file *module types*.

Sum output

The signals of the four HF channels are summed and this signal is available on an extra SMA connector on the back of the detector. It is the sum of the signals after the HF amplifiers, divided by four. This output is added for debugging purposes, since there is no other way to check the amplitude of the RF signal. This signal is not isolated from the signal ground. The output impedance is 50Ω.

Overload indication

There are 2 LEDs indicating an overload (signal out of range) for the DC and HF outputs. Each of the LEDs indicates that one or more of these channels are out of range. This means that the output voltage of one of the amplifiers is close to saturation. For the HF outputs of the amplifiers before subtraction are monitored. These LEDs are only visible when the cover is removed.

Noise

The pre-amplifier is a transconductance amplifier and can be looked upon as a simple resistor, converting current into voltage. In the old design an amplifier from Analog Modules is used, with a gain specified as $2kV/A$. This stage, together with the detector, largely determines the noise behavior of the whole system. The noise of the pre-amplifier consists out of three components, current noise I_N, the voltage noise E_N and the noise from the feed-back resistor. The voltage noise can be converted to a current component at the input of the amplifier, which is common design practice. If the transconductance amplifier is correctly compensated (no gain peaking) the bandwidth to determine the total noise is the -3 dB point F. The feedback resistor, determining the conversion factor, is R_F. C_D is the photo diode's capacitance. C_F is the feedback or compensation capacitor, which prevents gain peaking due to the detector's capacitance. Its value is small (~1 pF), and does not show up in the noise calculations.

First, one determines a stable (flat gain) configuration and then analyzes the noise behavior. For a maximally flat 2nd order Butterworth frequency response, a feedback pole is needed at $\frac{1}{2\pi R_F C_F} = \sqrt{\frac{GBP}{4\pi R_F C_D}}$. Where GBP is the gain-bandwidth product of the amplifier. So, choosing R_F determines C_F. This sets the system bandwidth a little lower than needed for stability, but gives a very predictable behavior. Also, this clarifies the need of a much larger GBP than one would expect at first glance.

The total noise density contribution of the amplifier, converted to the amplifier's input, can now be described as: $\overline{I_A} = \sqrt{\overline{I_N}^2 + \frac{4kT}{R_F} + \left(\frac{\overline{E_N}}{R_F}\right)^2 + \frac{(\overline{E_N} \cdot 2\pi F \cdot C_D)^2}{3}} \left[A/\sqrt{Hz} \right]$.

The signal from the detector is used in two ways: The so called DC outputs, which have a bandwidth of ~2 kHz. In this case this value can be filled in for F. Up to 100 Hz to 1 kHz the 1/f noise is dominant however.

The RF outputs run through a demodulator. There the signal is multiplied with a reference signal, thus converting the signal to a lower frequency. After that, it passes a 3 kHz low-pass filter (B). For the white noise, the first three terms in the equation, the noise is calculated by looking at a spectrum of two times 3 kHz around the center frequency of 6.26 MHz. Here we have to take into account both sidebands of the centre frequency, giving a total bandwidth of 6 kHz. In the last term however, the noise density is frequency dependent: $\overline{i_N} = \overline{E_N} \cdot 2\pi f \cdot C_D$. This, we have to integrate over the pass-band around the center (reference) frequency to get an RMS value.

$$\overline{i_N^{\,2}} = \int_{f_0-B}^{f_0+B} \left(\overline{E_N} \cdot 2\pi f \cdot C_D\right)^2 df \Rightarrow$$

$$i_{N,RMS} = \overline{E_N} \cdot 2\pi \cdot C_D \cdot \sqrt{\int_{f_0-B}^{f_0+B} f^2 df} = \overline{E_N} \cdot 2\pi \cdot C_D \cdot \sqrt{\frac{(f_0+B)^3 - (f_0-B)^3}{3}} =$$

$$\overline{E_N} \cdot 2\pi \cdot C_D \cdot \sqrt{\frac{4 f_0^{\,2} \cdot B + 2 f_0 \cdot B^2 + 2B^3}{3}} = \overline{E_N} \cdot 2\pi \cdot C_D \cdot \sqrt{3 f_0^{\,2} + B^2} \cdot \sqrt{\frac{2 \cdot B}{3}}$$

$$3 f_0^{\,2} >> B^2 \Rightarrow i_{N,RMS} \approx \overline{E_N} \cdot 2\pi \cdot C_D \cdot f_0 \cdot \sqrt{2B}$$

Or, referred to the input, looking at noise density: $\overline{i_N} \approx \overline{E_N} \cdot 2\pi \cdot C_D \cdot f_0 \cdot \sqrt{2} \left[A/\sqrt{Hz} \right]$

In the transconductance stage the OPA657 is used. The voltage noise of the OPA657 is 4.8 $nV/\sqrt{Hz}$ and the current noise is 1.3 $pA/\sqrt{Hz}$. For the OPA846 these figures are 1.2 $nV/\sqrt{Hz}$ and 2.8 $pA/\sqrt{Hz}$. Its gain-bandwidth product is much lower though.

Looking at the four terms of the equation above, converted to the output of this stage, we find:

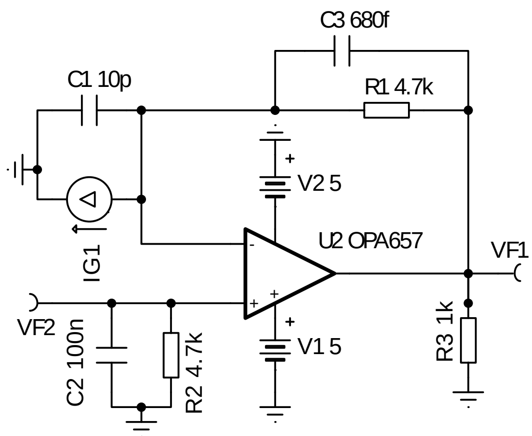
$$E_{o,RMS} = R_F \cdot \sqrt{\overline{I_N^{\,2}} + \frac{4kT}{R_F} + \left(\frac{\overline{E_N}}{R_F}\right)^2 + \frac{(\overline{E_N} \cdot 2\pi F \cdot C_D)^2}{3}} \left[V/\sqrt{Hz} \right] =$$

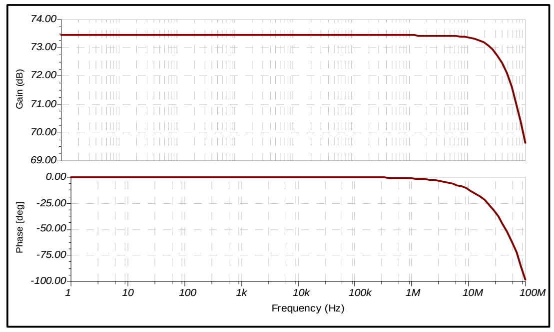
$$4.7 \cdot 10^3 \cdot \sqrt{\left(1.3 \cdot 10^{-12}\right)^2 + \frac{1.6 \cdot 10^{-20}}{4.7 \cdot 10^3} + \left(\frac{4.8 \cdot 10^{-9}}{4.7 \cdot 10^3}\right)^2 + \frac{\left(4.8 \cdot 10^{-9} \cdot 2\pi \cdot 3.3 \cdot 10^3 \cdot 10 \cdot 10^{-12}\right)^2}{3}} =$$

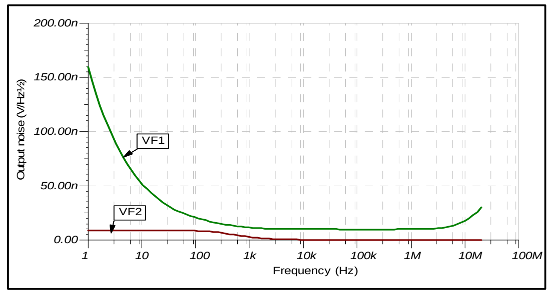
$$4.7 \cdot 10^3 \cdot \sqrt{1.7 \cdot 10^{-24} + 3.4 \cdot 10^{-24} + 1.0 \cdot 10^{-24} + 33 \cdot 10^{-32}} \left[V/\sqrt{Hz} \right]$$

For the white noise of the transconductance stage we thus get 12 $nV/\sqrt{Hz}$.
Around 6 MHz the contribution of last term changes to:
$4.8 \cdot 10^{-9} \cdot 2\pi \cdot 6 \cdot 10^6 \cdot 10 \cdot 10^{-12} / \sqrt{3} = 1.1 pA/\sqrt{Hz}$ or 5.2 $nV/\sqrt{Hz}$. This sums up to ~13 $nV/\sqrt{Hz}$. For the output noise this figure is amplified 3 times. There is the 3 dB loss in the filter and the six times amplification of the HF amplifier. The output signal is a combination of four channels. This sums up to ~70 $nV/\sqrt{Hz}$.

Below ~1 kHz the noise figure is dominated by the 1/f noise. This can only be determined by simulating the circuit, using correct Spice models. For these OpAmps these models are available within the TINA-TI simulator.
In the circuit below, the 10 pF in parallel with the current source represents the photo diode quadrant.



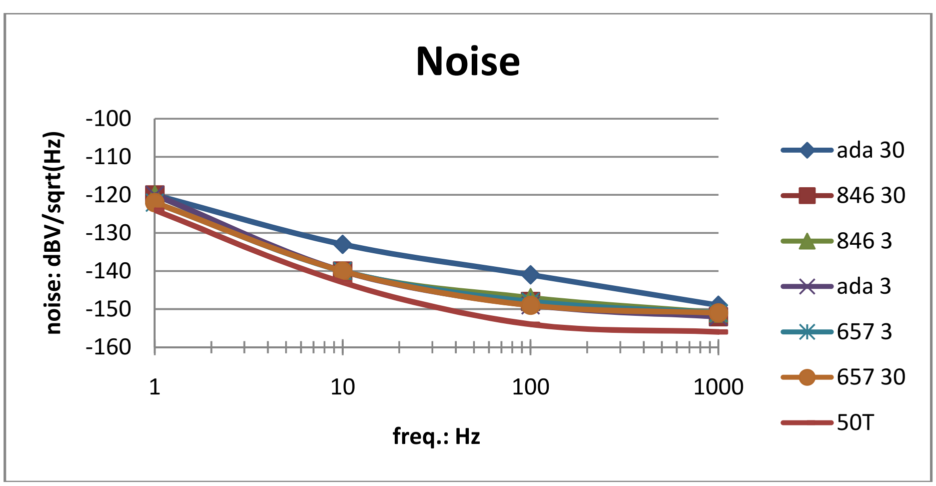




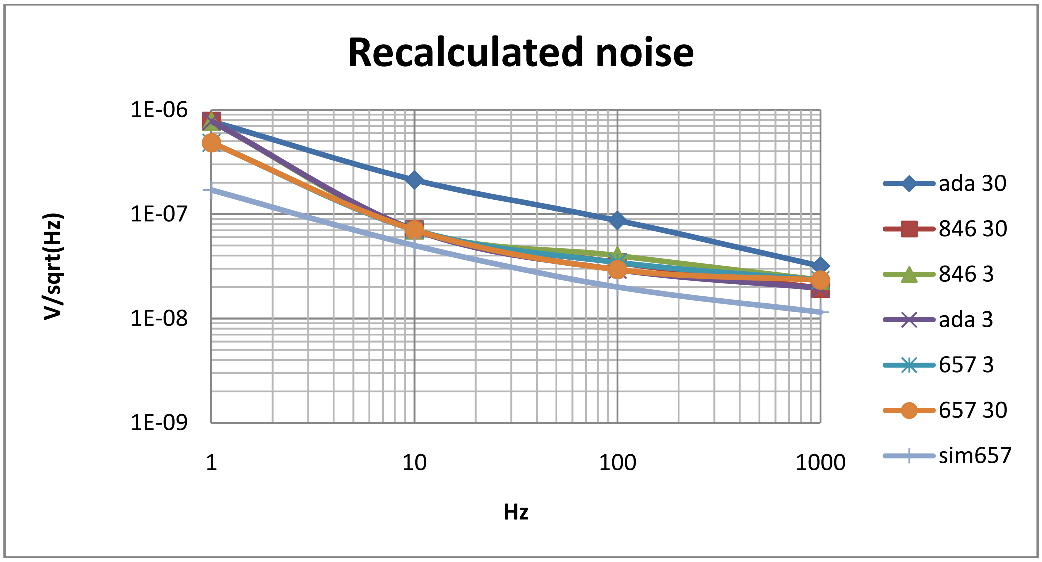
At low frequencies there is a small contribution (VF2 in the plot) from the current noise in the bias compensation resistor.

Measuring the LF noise

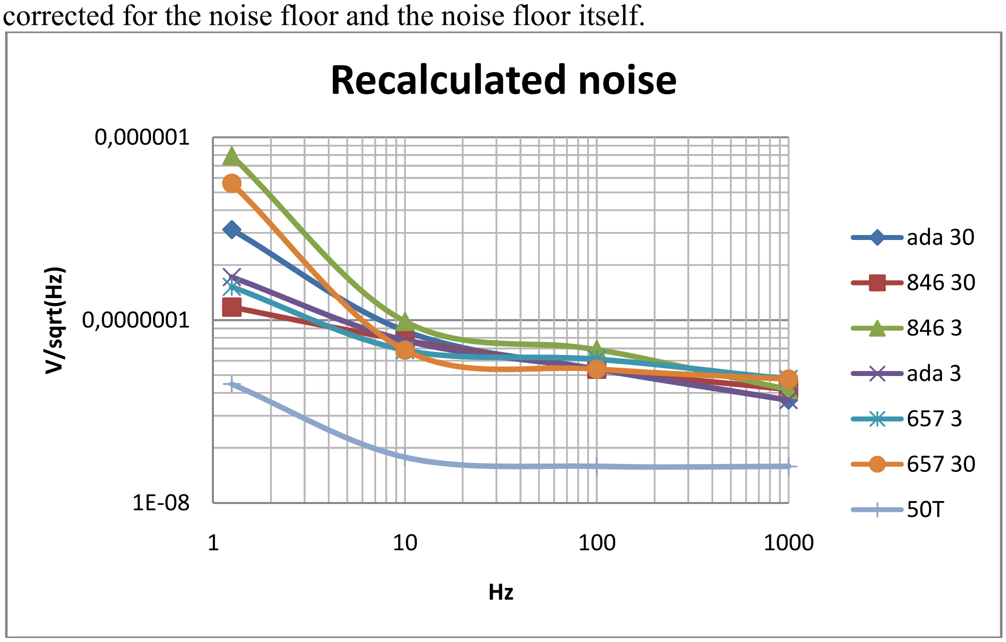
The voltage noise of the transconductance stage, combined with the detector capacitance is dominant at higher frequencies. Tests include an extremely low noise OpAmp, the ADA4899-1. Though having very promising (white) noise figures, it disappoints at low frequencies.
For the tests, six types of detector boards were build, with three types OpAmps: The OP657, the OPA846 and the ADA4899-1. Each type is used in the configurations 3 mW and 30 mW. The whitening filters are disabled.
For the measurement, an Advantest R9211 spectrum analyzer is used. We use an assembled module, including the photo diode. The signal from the transconductance stage is amplified by the whitening filters, with the whitening part set to 0 dB. Amplification is nearly a factor of two (1.9). This is lost again in the 50 Ω output, since the input impedance of the analyser was set to 50 Ω. So, the signal measured is approximately the same as the output of the transconductance amplifier.



The 50T line is a termination of 50 Ω instead of the module under test. It shows the measurement limit of the instrument. We can use this to correct the measurement. Since these are RMS values we can subtract the 50T values from the other measurement using $\overline{e_n} = \sqrt{\overline{e_{meas}}^2 - \overline{e_{50T}}^2}$



The simulation of the OPA657 is now included. Above 10 Hz the measurement agrees with the simulation within a factor of two.
In the previous plot one can see that, using just the termination, the line is not very flat. This probably indicates that there is more 1/f noise from the instrument itself then one would expect of an instrument that claims to measure down to 10 mHz.
The measurement was done again at Virgo, using a Ono Sokki analyser (range DC – 100 kHz). In this case the input of the analyser was set to high impedance. The plot contains the measurements, corrected for the noise floor and the noise floor itself.



One can see a much flatter noise floor, suggesting this instrument has a better behaviour at low frequencies. The noise voltage has increased, but not as much as one would expect, using the high input impedance. So all measurements are not far from the simulation values, except at very low frequencies. Also note that the measurement included the photo diode, which was biased. If its leakage current would be 10 nA, this would contribute $\sqrt{2 \cdot 1.6^{-19} \cdot 10^{-9}} = 5.6^{-14}\ {}^{A}/_{\sqrt{Hz}}$. For the 3 mW version, this gives $0.26\ nV/\sqrt{Hz}$ at the output of the transconductance amplifier. This can be neglected compared to the other values, but it is not a guaranteed figure and is affected by temperature, bias voltage and the photo current itself.

Measuring the HF noise

Measuring the HF noise of the horizontal and vertical outputs could only be done in a very indirect way. So this was done again at Virgo. For this measurement an Agilent 4395A was used (range 100 kHz – 500 MHz). The measurement was done around the module's centre frequency (6.26 MHz), from 5 to 7 MHz. Noise drops a bit on both ends, we measured the peak near the centre frequency.
The horizontal output gave $-138\ dBV/\sqrt{Hz}$, vertical $-137\ dBV/\sqrt{Hz}$. This is a little more than $100\ nV/\sqrt{Hz}$. Earlier $70\ nV/\sqrt{Hz}$ was estimated. So this is very close.