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AN4708
Application note

Signal conditioning for shock sensors

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Introduction

This application note deals with the analog signal conditioning circuit used in shock sensors that behave like piezoelectric sensors. The document explains how to condition a signal coming from a shock sensor and how to improve performance. It relates to all products of the TSX7 and TSX9 families.

Shock sensors can be used in a wide range of applications. They are mainly used in the consumer market for hard disk drive protection. To optimize reading, the reading head is placed extremely close to the platter. Shock sensors are able to detect acceleration which allows the driver to suspend a read or write operation, and to park the head, which protects the disk.

Shock sensors are also used in the automotive sector, for example, for security when a window is hit and broken. They can enable intelligent power management to maximize battery life for tire pressure monitoring system modules integrated in tire valves.

In addition, we find shock sensors in industrial applications, for example, to detect abnormal vibration in motors. In this case, the generated signal from the shock sensor is analog and it has to be amplified and filtered to be useful.

Contents

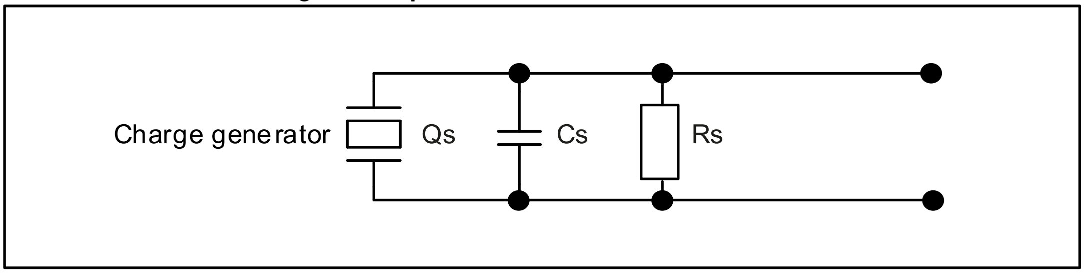
1 The piezoelectric effect

The piezoelectric effect was discovered at the end of the 19th century by the Currie brothers. They discovered that quartz changes its dimensions when it is placed in an electrical field and inversely, quartz is able to generate an electrical charge when pressure is applied.

Normally, piezoelectric crystals are electrically neutral. The atoms inside them may not be symmetrically arranged, but their electrical charges are perfectly balanced: a positive charge in one place cancels out a negative charge nearby. However, if you squeeze a piezoelectric crystal, you deform the structure, pushing some of the atoms closer together or further apart. This upsets the positive and negative balance and causes net electrical charges to appear. Conversely, a piezoelectric crystal deforms if a voltage is applied to it.

A shock sensor is generally modeled as a charge source with a shunt capacitor as shown in *Figure 1*. The induced charge, Qs, is linearly proportional to the applied force. The capacitance, Cs, is proportional to the surface area of film and is inversely proportional to the film thickness. The resistance, Rs, is generally very high (in the range GΩ).

Figure 1: Equivalent circuit of a shock sensor



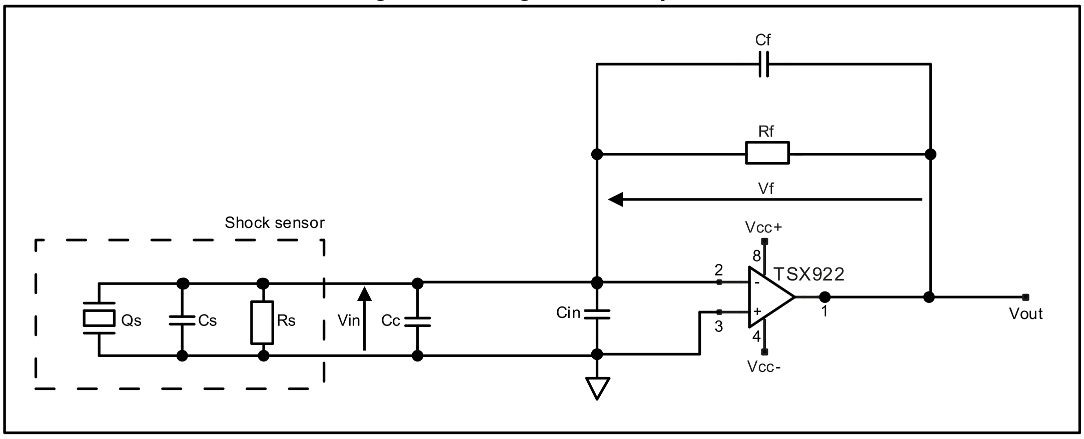
For signal conditioning of a shock sensor, there are two different approaches depending on the application. If the sensor is far from the electronics, it is better to use a charge amplifier (see *Figure 2*). If the sensor is close to the electronics, a simple voltage configuration might be sufficient (see *Figure 11*).

2 Charge amplifier configuration

Charge mode sensors are typically used when the electronics are connected far from the sensor. In this case, we can use the configuration shown in *Figure 2*.

The charge amplifier requires a low bias input current as it does not charge and discharge the gain capacitor, Cf, at high currents. Consequently, it is extremely important to choose a CMOS op amp such as the TSX922 which presents a very low input current, Iib, of 10pA @25 °C.

Figure 2: Charge mode amplifier



If any charge coming from the piezoelectric sensor "tries" to charge the capacitance of the sensor, the cable, or the input capacitance of the amplifier, a voltage is created between the input pin of the amplifier.

As the amplifier has a very high gain (90 dB), this voltage is immediately nulled by sourcing or pulling the same amount of charge through the feedback capacitance, Cf, and the resistance, Rf.

2.1 Gain

The input charge, Qs, is applied to the inverting input of the amplifier. It is distributed to the cable capacitance, Cc, the amplifier input capacitance, Cin, and the feedback capacitor, Cf.

$$Qs = QCc + QCin + QCf \qquad (1)$$

By considering that Q = CV we can write:

$$Qs = Vin(Cc + Cin) + Vf.Cf \qquad (2)$$

Where Vin is the differential voltage of the op amp and Vf is the voltage in the feedback loop.

Thanks to the large gain of the op amp (AVD), the same potential tends to exist between pin+ and pin-. Consequently, we can consider that Vin = 0 and simplify Equation (2) as follows:

$$Qs = Vf.Cf \qquad (3)$$

As Vout = -Vf:

$$Vout = -\frac{Qs}{Cf} \qquad (4)$$

From Equation (4) we can see that the charge amplifier gain is independent of the input capacitance, therefore the system sensitivity is unaffected by changes in input, cable length, or type. We also notice that the smaller the capacitance the bigger the gain (1/Cf).

2.2 Bandwidth

Referring to *Figure 2*, the function of the feedback resistor, Rf, is to provide DC stability to the circuit and to define the lower frequency limit of the amplifier. At first, it seems that the Rf resistor, which is associated with the gain capacitance, Cf, acts as a low-pass filter. This would be true for a transimpedance configuration where the current is converted into a voltage thanks to Rf. But in this case, the charge is converted into a current thanks to the Cf capacitance. To better understand this concept, we need to look in more detail at the closed-loop transfer function.

$$Vout = -\frac{j\omega . Qs . Rf}{1 + j\omega Rf . Cf} \qquad (5)$$

In Equation (5) above, we can see that the Rf resistor, associated with the gain capacitance, Cf, creates a high-pass filter, with a cut off frequency of -3 dB.

In Equation (6) we can see that the Rf resistor must be as high as possible to keep the pole low.

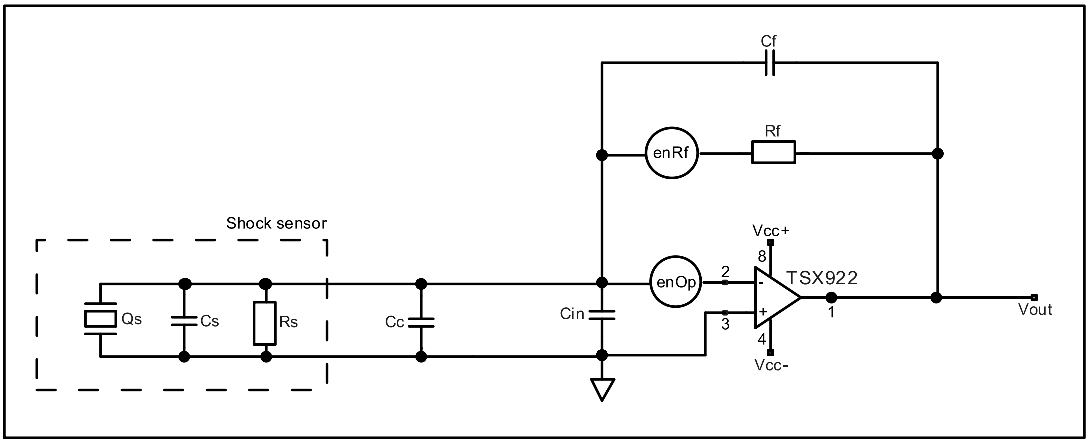
$$fhp = \frac{1}{2\pi . Rf . Cf} \qquad (6)$$

2.3 Noise

From Equations (4) and (6) we can see that for a piezoelectric sensor working as a shock sensor, the value of the feedback capacitance, Cf, should be decreased to increase the gain. But, in this regard, it is also important to increase the Rf resistor to keep the high-pass filter frequency as low as possible. Unfortunately, increasing the resistance introduces noise and increasing it too much may cause problems. To deal with this, we need to look at the spectral noise density.

Figure 3 shows the charge mode amplifier with the main noise sources: "enOp" is the voltage noise generated by the op amp and "enRf" is the voltage noise generated by the feedback resistor.

Figure 3: Charge mode amplifier with noise source



The output noise spectral density can be expressed as shown in Equation (7). Note that the impact of Rs noise is very low compared to the impact of Rf noise.

$$enTot = \sqrt{enOp^2 . \left|1 + \frac{j\omega Rf\,(Cs + Cc + Cin)}{1 + j\omega Rf . Cf}\right|^2 + enRf^2 . \left|\frac{1}{1 + j\omega Rf . Cf}\right|^2} \qquad (7)$$

Generally, the frequency range (when amplification is proportional to 1/Cf) is large enough to consider that the noise, before the high-pass cut off frequency (fHp), has a very low contribution to overall noise. Consequently, Equation (7) can be simplified, especially if we take into account that the sensor is used above the high-pass filter frequency as shown in Equation (8).

$$enTot = \sqrt{enOp^2 . \left|1 + \frac{(Cs + Cc + Cin)}{Cf}\right|^2 + enRf^2 . \left|\frac{1}{1 + j\omega Rf . Cf}\right|^2} \qquad (8)$$

Next, let us consider that the Cf capacitance value is decreased and at the same time the Rf resistance value is increased to keep the same high-pass filter cut off frequency:

$$\left(\frac{1}{2\pi . Rf . Cf} = cst\right)$$

From Equation (8), we can see that the noise related to the op amp (the first term of the equation) increases linearly with Rf (considering that Cf << Cs + Cc + Cin). Simultaneously, the noise related to the feedback resistance (the second term of the equation) increases, following the thermal noise of the resistance, expressed as enRF = $\sqrt{4KTRf}$, so as $\sqrt{Rf}$ K is the Boltzmann's constant, $1.38.10^{-23} JK^{-1}$, and T is the temperature in °K. We can see that "enTot" increases at a lower rate than the Rf.

The gain of the circuit is proportional to 1/Cf, so if we keep the fHp constant it will be proportional to the Rf.

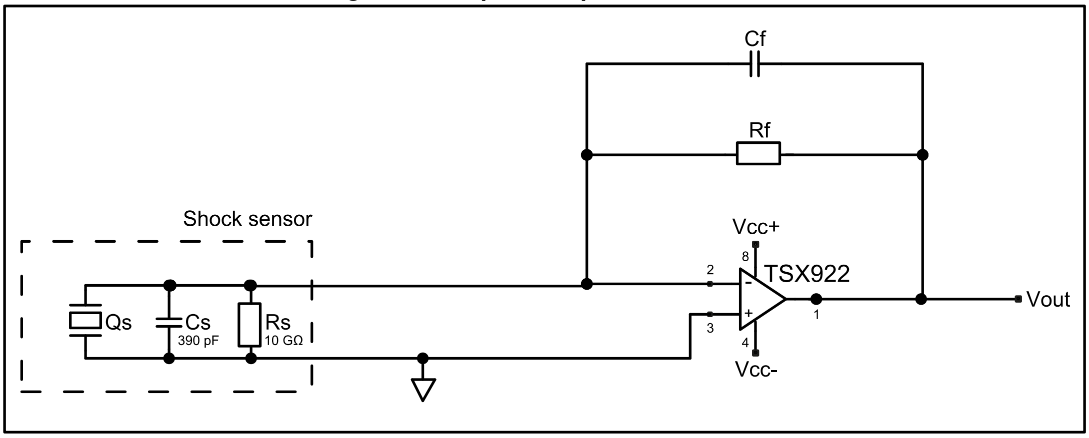
Consequently, increasing the Rf improves the signal-to-noise ratio (SNR). Effectively, doubling the Rf resistance allows the Cf capacitance to be divided in two (while still keeping the same high-pass filter). The gain and the output signal are doubled as a result. Note that in Equation (8) the noise is multiplied by a value lower than two.

Also, note that the Rf resistance value cannot be increased infinitely because very high resistance values are extremely difficult to implement on a PCB. Furthermore, the Rf combined with the input current, Iib, of the op amp generates a non-negligible DC offset on the output of the op amp.

2.4 Stability

Shock sensors are generally very capacitive and the Rf and Cf must be chosen carefully as they have a direct impact on the stability of the system (see *Figure 4*).

Figure 4: Simplified equivalent circuit



Equation (9) below relates to the open-loop transfer function of the system where "A" is the open-loop transfer function of the op amp.

$$-A\frac{Rs}{Rs+Rf} * \frac{1+j\omega(Cf.Rf)}{1+j\omega\left(\frac{Rs.Rf}{Rs+Rf}.(Cf+Cs)\right)} \qquad (9)$$

Equation (9) produces one pole which is shown below in Equation (10).

$$fp = \frac{1}{2\pi.\frac{Rf.Rs}{Rf+Rs}.(Cf+Cs)} \qquad (10)$$

Equation (9) also produces one zero which is shown in Equation (11).

$$fz = \frac{1}{2\pi.Rf.Cf} \qquad (11)$$

We also need to consider the low-frequency pole of the open-loop transfer function of the op amp as shown in Equation (12).

$$fop = \frac{GBP}{Avd} \qquad (12)$$

The bode diagram of this system can be plotted as shown in *Figure 5*. We can consider Rs >> Rf, so

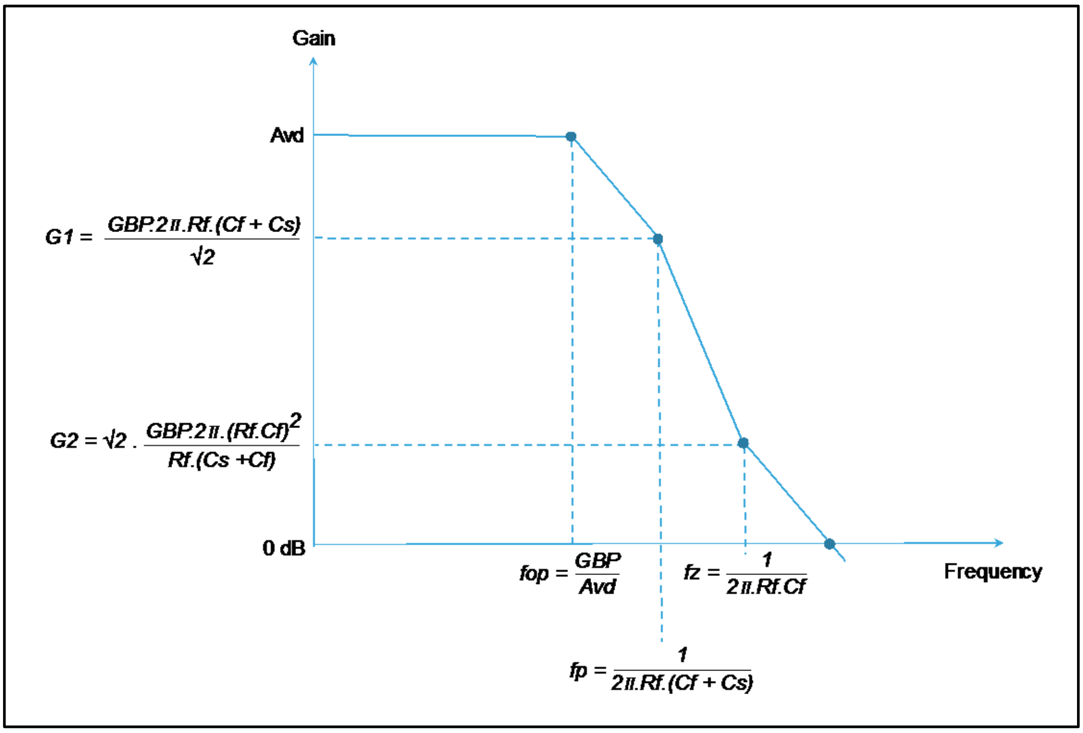
$$\frac{Rf.Rs}{Rf+Rs} \approx Rf$$

and

$$fp \approx \frac{1}{2\pi.Rf.(Cf+Cs)}$$

so, fz > fp.

Figure 5: Bode diagram of the open-loop transfer function of an application using a shock sensor



To guaranty the stability of the system, the bode diagram must cross the X-axis with a slope of -20 dB/decade.

To ensure stability in *Figure 5*, the gain at frequency, fz, must be higher than 1.

Consequently, we get Equation (13).

$$\sqrt{2}.\frac{GBP.2\pi.(Rf.Cf)^2}{Rf.(Cs+Cf)} > 1 \qquad (13)$$

From Equation (13), we can deduce the second order Equation (14).

$$\sqrt{2}.GBP.2\pi.Rf.Cf^2 - Cf - Cs > 0 \qquad (14)$$

Next, we look at an example of an application where abnormal vibrations in motors used in industrial electrical automation systems were detected. In the example, the motor is rotating at a speed of 500 Hz and the maximum shock that must be detected is 50 G.

We consider a shock sensor with the following characteristics:

- Cs = 390 pF
- Rs = 10 GΩ
- Sensitivity = 0.35 pC/G

First, we fix the gain thanks to a feedback capacitance, Cf = 100 pF. In the frequency range of interest, we obtain:

$$Vout = -\frac{Qs}{Cf} = -\frac{0.35}{100} = -3.5mV/G$$

To keep the system stable:

$$Rf > \frac{Cf + Cs}{\sqrt{2}.GBP.2\pi.Cf^2}$$

So, Rf > 550 Ω.

We set Rf to 10 MΩ to obtain a high-pass filter with a cut off frequency as low as possible (160 Hz) which prevents filtering the 500 Hz frequency.

2.5 Filtering

The shock sensor, which is considered as a piezoelectric element, has its own resonance frequency. In general, this frequency has to be filtered. In the example below, the shock sensor has a resonance frequency around 28 kHz.

Using the application schematic shown in *Figure 6*, we see the frequency response in *Figure 7* which shows the effect of the high-pass filter and the resonance peak.

To obtain the frequency response in *Figure 7*, the shock sensor has been used in series with the signal generator.

The high-pass filter is set at 160 Hz (see Equation (8)).

Figure 6: Application schematic

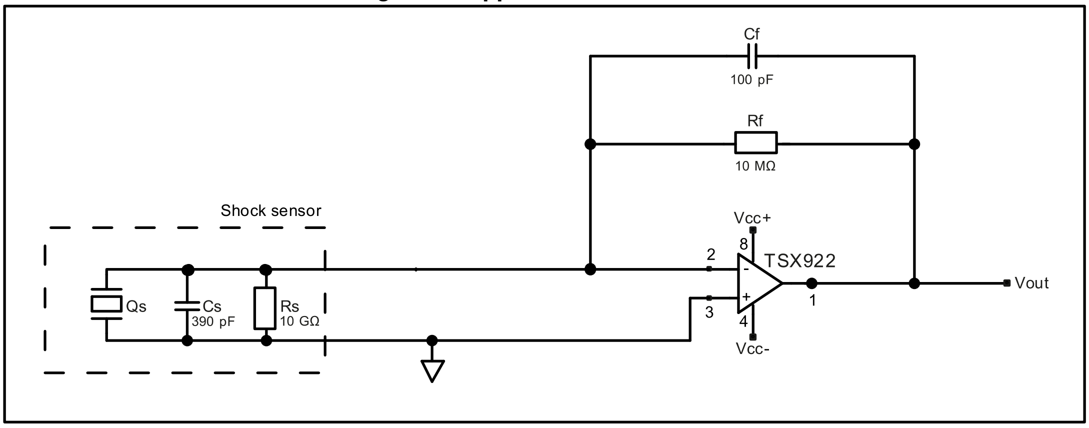
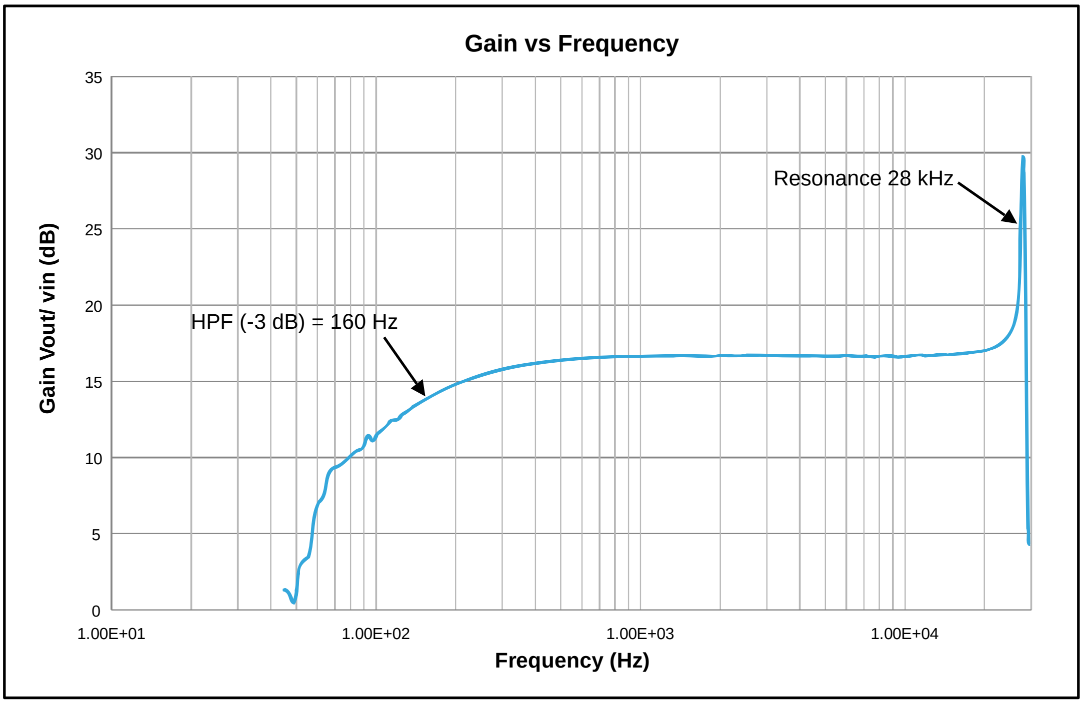


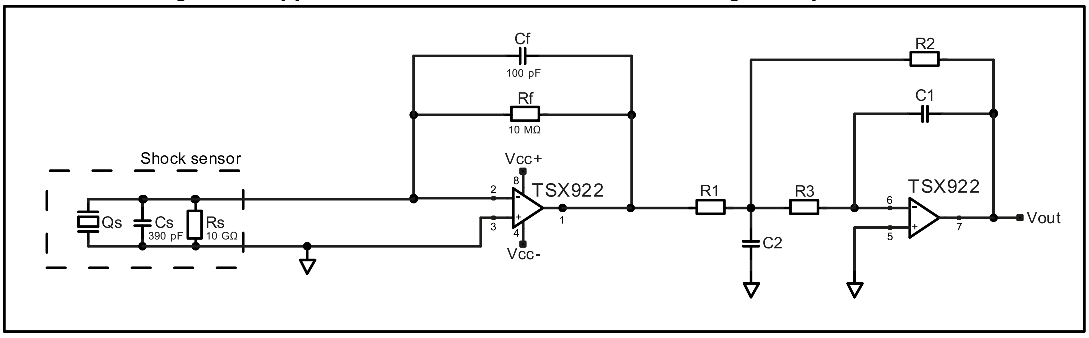
Figure 7: Typical amplitude response vs frequency



Adding a second stage low-pass filter, attenuates the resonance peak and avoids any negative effects in the application.

We chose a multiple feedback topology, Butterworth filter to provide the maximum pass band flatness and to obtain a quality factor of Q = 0.707. This schematic is shown in *Figure 8*.

Figure 8: Application schematic with a second stage low-pass filter



In *Figure 8*, the voltage signal of the first stage is generally small. We use the second-stage low-pass filter to add more gain to the whole schematic. The low-frequency gain of the second stage is given by Equation (15).

$$A0 = -\frac{R2}{R1} \qquad (15)$$

The corner frequency of a multiple feedback filter is given by Equation (16).

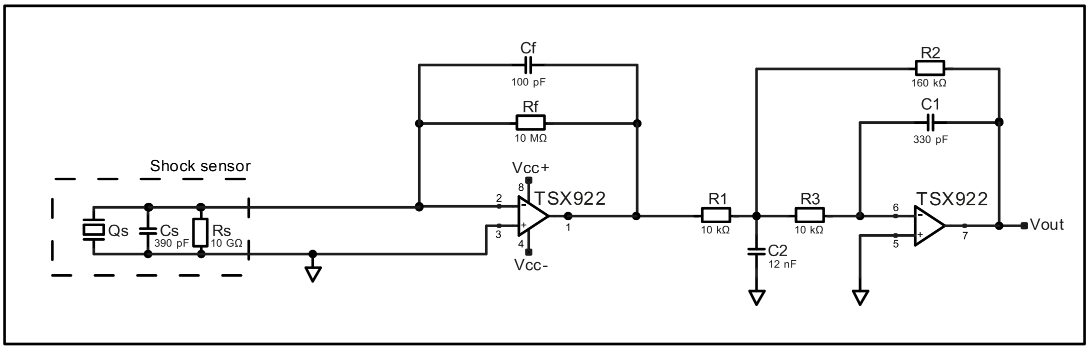
$$\omega 0 = \frac{1}{\sqrt{R3.R2.C1.C2}} \qquad (16)$$

In the application example of *Figure 8*, the resonance peak must be filtered at 28 kHz. So, we implement a low-pass filter with a cut off frequency of 2 kHz.

For filter calculations, use our filter tool on eDesignSuit which is available on the ST web site at: *http://www.st.com/web/en/support/eDesign.Suite.html*

Next, we choose the component values according to *Figure 9*.

Figure 9: Low-pass filter with a cut off frequency of 2 kHz



The gain of the second stage is now:

$$A0 = -\frac{160k\Omega}{10k\Omega} = -16\,V/V$$

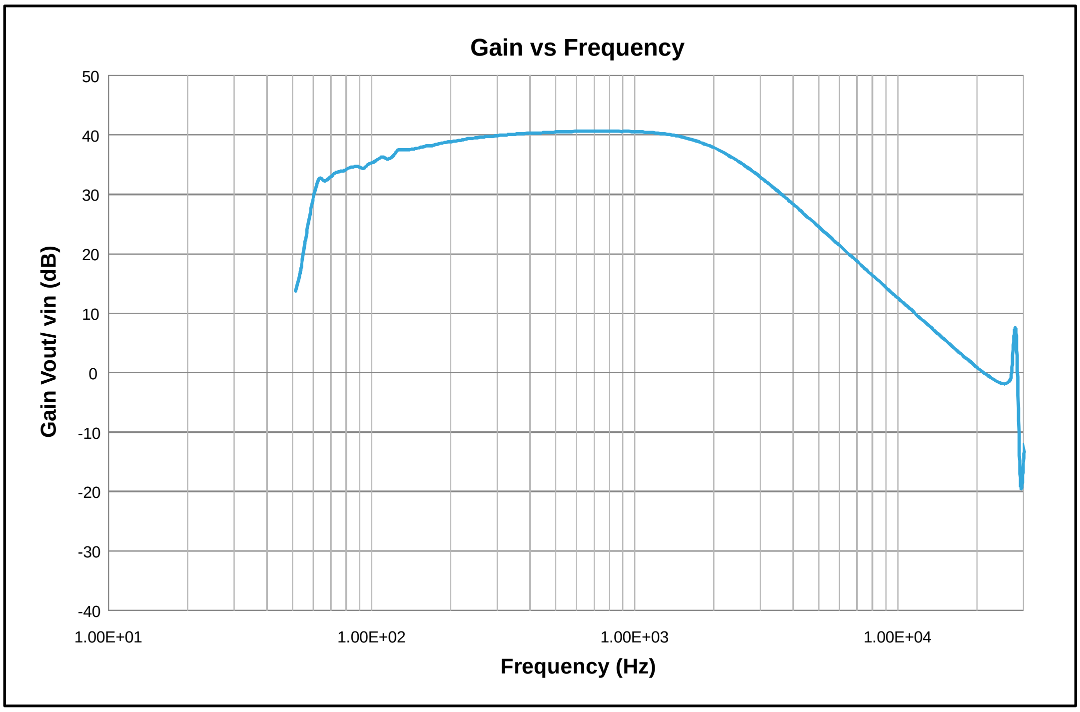
The cut off frequency (-3 dB) is:

$$f0 = \frac{1}{2\pi\sqrt{10k\Omega.\,16k\Omega.\,330pF\,.\,12nF}} = 2\ kHz$$

Figure 10 shows how the resonance peak has largely been attenuated. In addition, the gain for the whole system is 40 dB, whereas previously it was 17 dB.

To obtain the frequency response in *Figure 10*, the shock sensor has been used in series with the signal generator.

Figure 10: Typical amplitude response vs frequency with a low-pass filter



If the shock sensor is used in the range of 160 Hz to 2 kHz, we obtain the following:

$$Vout = \left(-\frac{Qs}{Cf}\right) * \left(-\frac{R2}{R1}\right) = \left(-\frac{0.35pC/G}{100pF}\right) * \left(-\frac{160k\Omega}{10k\Omega}\right) = 56mV/G$$

To complete the calculations, we can also look at the DC error voltage issued from such a configuration. A more detailed equation can be written as shown in Equation (17).

$$Vout_{DC} = \pm(iiB * Rf \pm Vio) * \left(-\frac{R2}{R1}\right) \pm Vio * \left(1 + \frac{R2}{R1}\right) \qquad (17)$$

Replacing Equation (17) with real values, we achieve:

$$Vout\_DC = \pm(100pA * 10M\Omega \pm 5mV) * \left(-\frac{160k\Omega}{10k\Omega}\right) \pm 5mV * \left(1 + \frac{160k}{10k\Omega}\right)$$

In the worst case, this gives an output swing above and below -181 mV.

The op amp is used with dual supply to avoid output saturation.

2.6 Time constant

An important advantage of using a charge amplifier is that the time constant is only due to the feedback resistance, Rf, and capacitance, Cf, and not from the piezoelectric or connecting cables. Consequently, G variation is detected more quickly.

The time constant is given by Equation (18).

$$\tau = Rf.Cf = 10M\Omega * 100pF = 1ms \qquad (18)$$

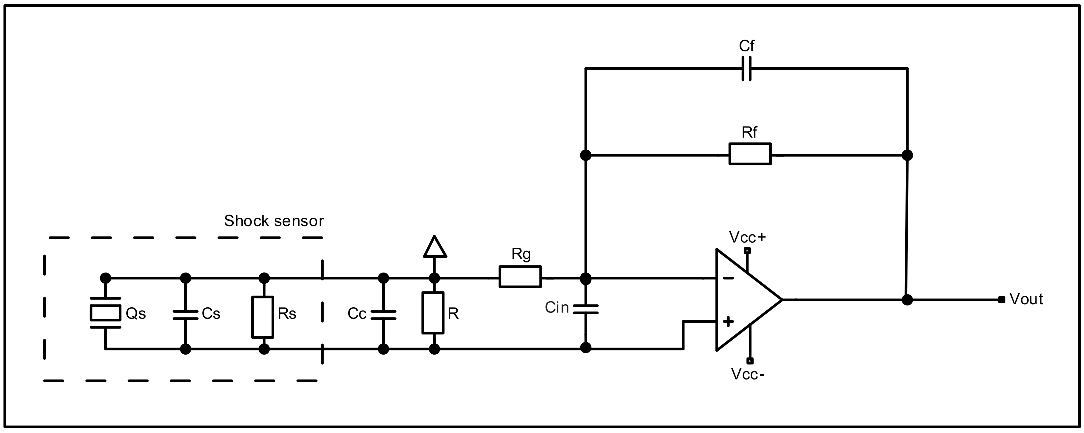
Using a charge amplifier configuration allows a transfer function that is independent of the capacitance of the shock sensor and the connecting cable. In turn, the connecting cable between the shock sensor and the signal conditioning can be easily changed without affecting the output voltage result.

Charge mode configuration also permits the use of a long cable between the electronics and the shock sensor. The sensor can be placed in a high temperature environment, generally higher than 125 °C (without exceeding the Curie temperature) and have a deported electronic in ambient temperature.

3 Voltage amplifier configuration

For the voltage mode amplifier the induced voltage is presented to the high impedance non-inverting input and then amplified by the op amp. The main advantage of voltage mode configuration is that the gain is set accurately with resistors rather than with a small capacitor (see *Figure 11*).

Figure 11: Voltage amplifier configuration



In a frequency range, all the charges generated by the sensor are transferred into Cs and Cc. The op amp amplifies this voltage as shown in Equation (19).

$$Vout = \frac{Qs}{Cs + Cc} * \left(1 + \frac{Rf}{Rg}\right) \qquad (19)$$

As the gain is related to the amount of capacitance seen by the sensor, the shock sensor must be connected as close as possible to the op amp in this configuration. This is because the parasitic capacitance of the cable, Cc, affects the actual gain (and the longer the cable, the higher this capacitance). It is important to add a resistance, R, so that the DC correctly biases the op amp.

One of the main advantages of the voltage amplifier configuration is its simplicity. To better understand the behavior of this configuration let us develop the transfer function over the frequency range as shown in Equation (20).

$$Vout = \frac{\frac{Rs.R}{Rs + R}.j\omega Q}{1 + j\omega\frac{Rs.R}{Rs + R}.(Cc + Cs)} * \left[\frac{\frac{Rf}{Rg}}{1 + j\omega Cf.Rf} + 1\right] \qquad (20)$$

From Equation (20), we can see that the voltage mode amplifier naturally provides a passband frequency response. Effectively, from Equation (20) we can see a high-pass filter with a cut of frequency, fHp, defined as shown in Equation (21).

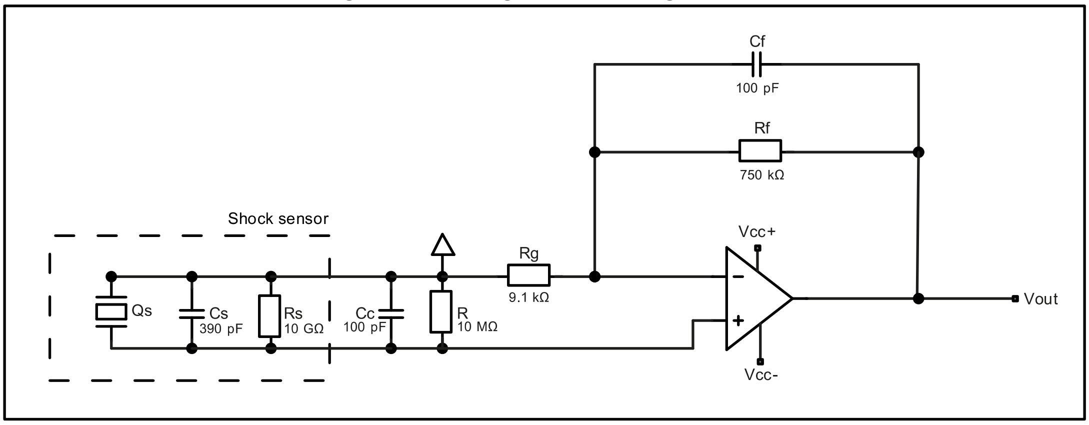
$$fHp = \frac{1}{2\pi.\frac{Rs.R}{Rs + R}.(Cs + Cc)} \qquad (21)$$

A low-pass filter with a cut of frequency, fLp, is defined by Equation (22).

$$fLp = \frac{1}{2\pi.Rf.Cf} \qquad (22)$$

Then, let us take the same example as the charge mode configuration (see *Figure 12*). In this case, abnormal vibrations in the motors are detected, Qs = 0.35 pC/G. Consequently, we must take the length of the cable into consideration, Cc = 100 pF for a cable of 1 m. To work in low frequency, the resistance, R, must be as high as possible, R = 10 MΩ.

Figure 12: Charge mode configuration

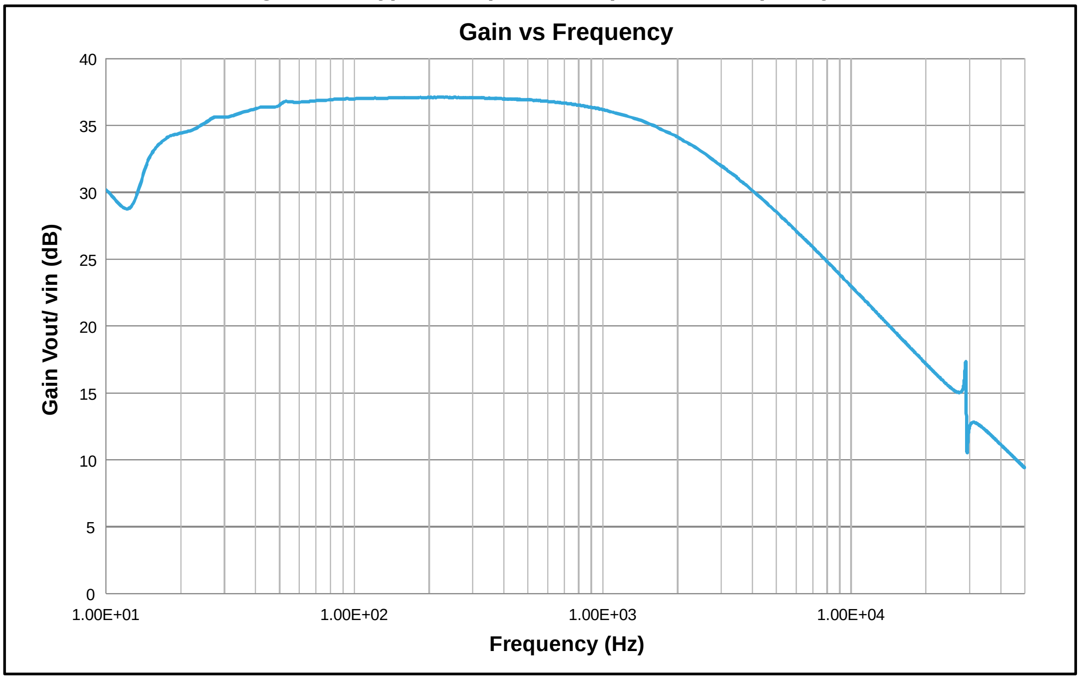


The high-pass filter is set at 32.5 Hz (see Equation (21)).

Next, let us choose a low-pass filter with a cut off frequency set at 2 kHz and Cf = 100 pF. From Equation (22), we can deduce that Rf = 750 kΩ. To be in the same range as the application in charge mode configuration, we need a gain of 56 mV/G. We deduce that Rg = 9.7 kΩ.

Figure 13 shows that with such a configuration, the resonance peak has been attenuated and that the gain for the whole system is 37 dB.

Figure 13: Typical amplitude response vs frequency



To obtain the frequency response in *Figure 13*, the shock sensor has been used in series with the signal generator.

In voltage mode amplifier, it is important to consider the error linked to the DC parameter of the amplifier, especially the input current bias (Iib) and the input offset voltage (Vio). They introduce a DC offset on the output defined in Equation (19).

Equation (19) can be developed with the DC parameters as shown in Equation (23).

$$Vout_{DC} = iiB * R * \left(1 + \frac{Rf}{Rg}\right) \pm Vio * \left(1 + \frac{Rf}{Rg}\right) \qquad (23)$$

Consequently, it is better to choose an amplifier with a low input voltage offset (Vio) and low input bias current (Iib). The TSX711A is a good part in this respect as it presents a maximum Vio of 350 µV over temperature. However, note that the Iib of the TSX711A cannot exceed 200 pA over temperature.

With the above conditions and using Equation (23), the maximum Vout DC voltage is:

$$Vout = 200pA * 10M\Omega * (1 + 82.4) \pm 350\mu V * (1 + 82.4)$$

Vout = 196 mV.

Therefore, the output swings above and below a DC voltage located somewhere between ±196 mV and the amplification in the frequency range of interest is:

$$Vout = \frac{0.35pC/G}{100pF + 390pF} * (1 + 82.4) = 60mV/G$$

We can see that even by choosing a precision amplifier, the voltage gain configuration is more sensitive to the DC component than the charge amplifier configuration. These calculations do not take into account the precision of the resistance used (generally 1 %) or their variation over temperature.

A solution to remove the DC output offset is to add a serial coupling capacitance on the output. But in this case, depending on the output load, a low-pass filter is created which directly impacts on the low-frequency response.

3.1 Time constant

Contrary to the charge amplifier configuration, the time constant of a voltage gain configuration mainly depends on the piezoelectric element and the connecting cable.

The time constant of such a system is defined in Equation (24).

$$\tau = R.(Cs + Cc) = 10M\Omega * 490pF = 4.9ms \qquad (24)$$

A voltage gain configuration is suitable for low-frequency response systems.

4 Conclusion

A piezoelectric accelerometer as a shock sensor can be used with either a charge mode configuration or a voltage mode configuration.

Charge amplifiers sense the charge coming from the shock sensor. They transform this charge thanks to the feedback capacitance. As the charge is the parameter sensed, the system is unaffected by the length of the cable. This is the big advantage of the charge amplifier configuration over the voltage amplifier configuration i.e. the couple shock sensor and amplifier can be calibrated in the laboratory with any convenient length of cable and then used directly in any kind of application. The frequency response of the charge amplifier is generally fast and depends only on the frequency response characteristic of the amplifier. With a charge amplifier, it is often necessary to add:

- a second stage to add gain
- a low-pass filter to remove the resonant frequency of the shock sensor

The voltage amplifier configuration is a simpler architecture. The gain and the band-pass frequency response can be set with only one op amp. This configuration generally exhibits quite a high time constant. Contrary to the charge amplifier configuration, sensitivity is reduced if long interconnecting cables are used between the shock sensor and the amplifier.

Depending on the application it is generally advantageous to have an amplifier with a large GPB. In this respect, the TSX922 is a good choice for the piezoelectric application using a charge amplifier configuration. When a voltage amplifier configuration is chosen, the TSX711A is a good choice of device thanks to its low offset input voltage.

5 Revision history

Table 1: Document revision history

Date	Revision	Changes
06-Aug-2015	1	Initial release