vibro-meter

MEGGITT

APPLICATION NOTE:

Cable length and attenuation in frequency for vibration measuring chains

CONTENTS

1

INTRODUCTION

Protection, control, and condition monitoring systems gather data from all connected sensors and data sources into a system, from which decisions are taken, and at the same time, analysis and predictive maintenance is performed on the monitored assets.

When sensors are located substantial distances from the central monitoring system and wireless communications are not suitable, the length of such interconnecting wiring must be carefully considered to ensure that the sensor's signal content is preserved[1].

There are many reasons why it may be necessary to incur long wiring distances between a monitor and its connected sensors. For example, harsh conditions at the sensor location may not allow the monitoring system to be co-located due to temperature, humidity, steam, dust, moisture, or aggressive chemical or flammable atmospheres present. Or, a local mounting location may be acceptable environmentally, but highly inconvenient or expensive for personnel to access. At the same time, local modules cannot achieve the goals of a centrally managed system, which can take decisions based on a large data set, and synchronised in real-time with other information at plant level.

Typical examples of such applications can include, but are not limited to:

- Refineries where assets (such as pumps in a tank farm) are a long distance from control rooms or enclosed buildings suitable for monitors, outside the explosive environment zone.
- Nuclear installations where the monitor is desired in a control room and the assets are in containment areas, or in remote pumping stations;
- Underwater measurement points such as pumps in canals, offshore locations, or other inaccessible locations.

Preservation of signal content is a particularly acute concern when dynamic signals – such as vibration and acoustic pressure – are being monitored. These signals frequently possess high-frequency content that is necessary for the integrity of the measurement and for proper diagnostics. Long cables are also necessary when real-time synchronisation is required with other parameters at plant level. Whatever the reason, applications with a need for long wiring runs occur often enough in practice to warrant guidance on this important topic. This Application Note has been authored specifically to address this need, explaining the calculations and considerations involved in computing cable length *versus* frequency response* and in contrasting the performance of different signal transmission schemes that can be employed to clearly highlight their differences. Several of these transmission schemes are unique to vibro-meter products and afford much longer cable lengths while preserving signal integrity than would otherwise be feasible. Special attention is thus given herein to explaining these schemes as they are not discussed in most other literature on the topic of cable length limitations. For safety applications of IPC707 electronics, under SIL-1 or SIL-2 certifications, the installation shall follow criteria on the system power supply capability, and on total capacitance [7], of which of course the cable acceptable length highly depends, as explained in §2.4 and §3.

[1] Signal content (i.e., frequency response) of field wiring is not the only concern when determining the maximum length of field wiring, but is the sole focus of this Application Note. A second constraint of field wiring length is the total amount of stored energy in the conductors when the installation is in a hazardous area and intrinsically safe practices will be used to achieve safety compliance. This topic is addressed in a separate Application Note. For installations in hazardous areas, both criteria must be assessed and the most restrictive wiring length then selected.

1. General considerations

1.1 Cable selection

Cable characteristics are of major importance for long transmission distances. Desirable cable characteristics for most field wiring applications are low values for distributed resistance, capacitance, and inductance. So-called "cable quality" generally represents cable with low values for these parameters, but at increased cost compared to lower-quality cable with higher values. However, depending on the application, the impact of these parameters will vary and may or may not be of concern. Thus, the cable quality that is necessary will be a function of the application.

For quasi-static signals, the primary cable parameter of concern is the distributed resistance (R_{cable} – typically expressed in *Ω/m* or *Ω/km*). However, for dynamic signals such as vibration and acoustics, both the internal resistance and the internal capacitance (C_{cable} – typically expressed in pF/m) are of concern. This is because the cable resistance and capacitance form a complex low-pass filter and signal content with frequencies around F_C (the filter's so-called cut-off frequency at -3dB) and beyond will be attenuated. Refer to §2.4 for additional information.

A cable's frequency response is characterized primarily by its internal capacitance, so this becomes the most important parameter when assessing cable length against its desired passband. In contrast, the cable's distributed inductance (L_{cable}, - typically expressed in μH/m) is secondary in such calculations and is indeed usually so small that it can be safely omitted when it comes to frequency response. However, the reader must be aware that although cable inductance is not discussed at length in this Application Note, and is rarely of concern for frequency response considerations, it does not mean the parameter is never of concern. Quite to the contrary, it can indeed play a substantial or even dominant role in hazardous area calculations.

Regardless of the application, vibro-meter universally recommends the following construction characteristics for field wiring cables when dynamic signals are transmitted:

- **Twisted wires**, by groups (whether pairs or triads),
- **Individually shielded groups**; in addition, the inclusion of an overall shield for multi-core cables is strongly recommended.

Both of these construction characteristics are important to protect transmitted signals from disturbances due to magnetic and electric fields. Twisting the conductors contributes to mostly magnetic field immunity (and cross-talk suppression *within* groups) while shielding the conductor groups contributes to mostly electric field immunity (and suppresses cross-talk *across* groups).

Multi-core cables are not recommended for sending several different signals, if within a single overall shield, or if twisted all together.

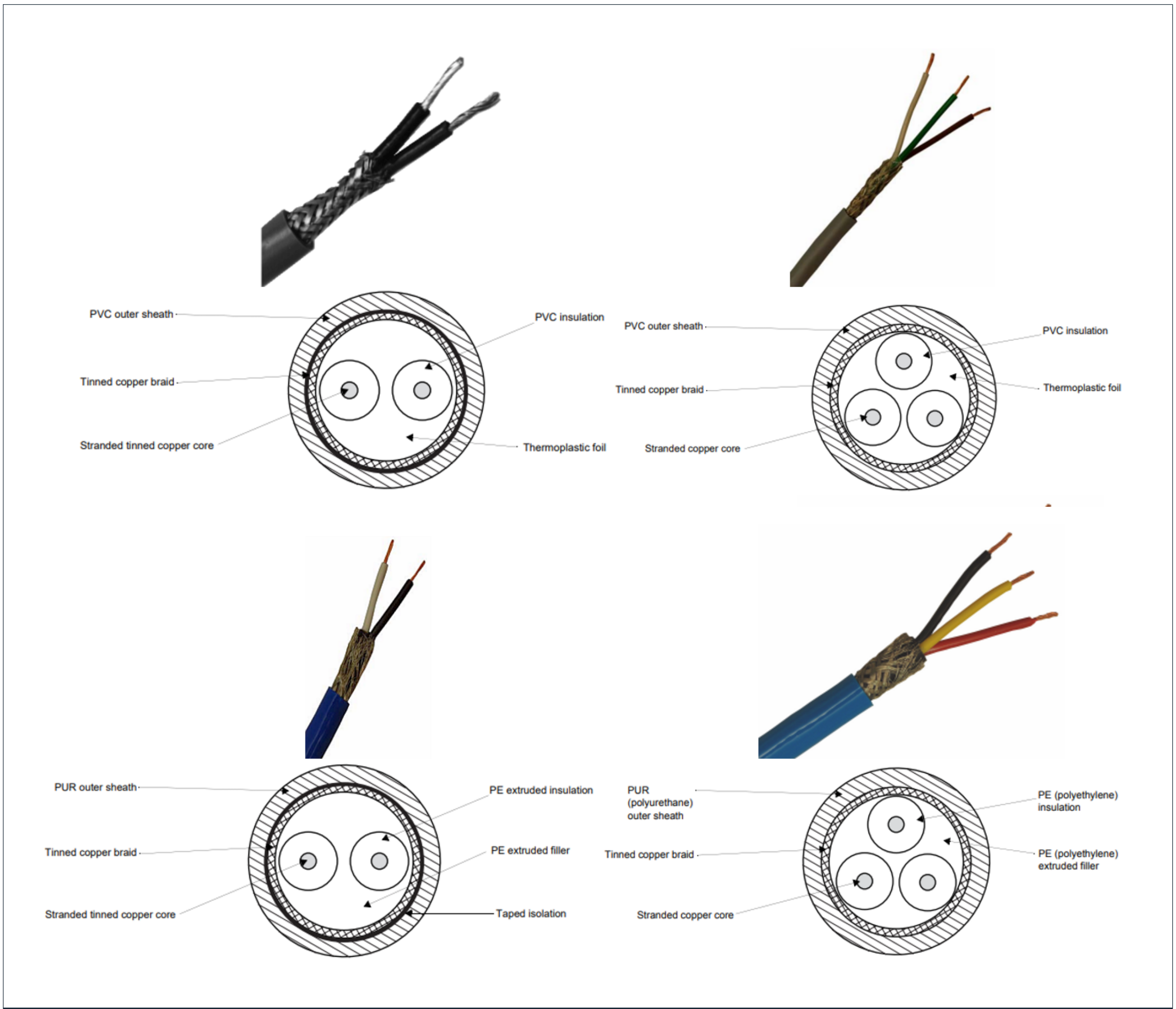

Figure 1: Examples of 4 vibro-meter tranmission cables suitable for long distance transmission. K209 and K210 cables (left) are used for 2-conductor sensors; and K309 and K310 cable (right) are used for 3-conductor transducers. Kx10 (below, blue), are of better quality than Kx09 (upper) and are approved for Exi applications.

For long transmission cables, price becomes an issue, leading to a compromise between cable cost and cable quality. Later in this Note, we will provide examples using several different cables in the vibro-meter catalogue representing different costs and qualities.

It is worth noting that the term “Low-noise” when referring to cables is not the same as the characteristics we have discussed thus far. Instead, it refers to another issue: the low-noise transfer of charge outputs (in picocoulombs) from piezoelectric sensors. These low-noise cables exhibit a particularly low tribo-electric effect and should always be used between the piezo-electric sensor and its corresponding charge amplifier even though the distances involved are relatively short. While such cables must also have low capacitance, low resistance, and low inductance, they must also possess a low tribo-electric effect – a quality not generally imposed upon the field wiring. The K221 cable used in vibro-meter’s EC222 cable assembly is an example of a low-noise cable intended for use between the piezoelectric device and its corresponding charge amplifier.

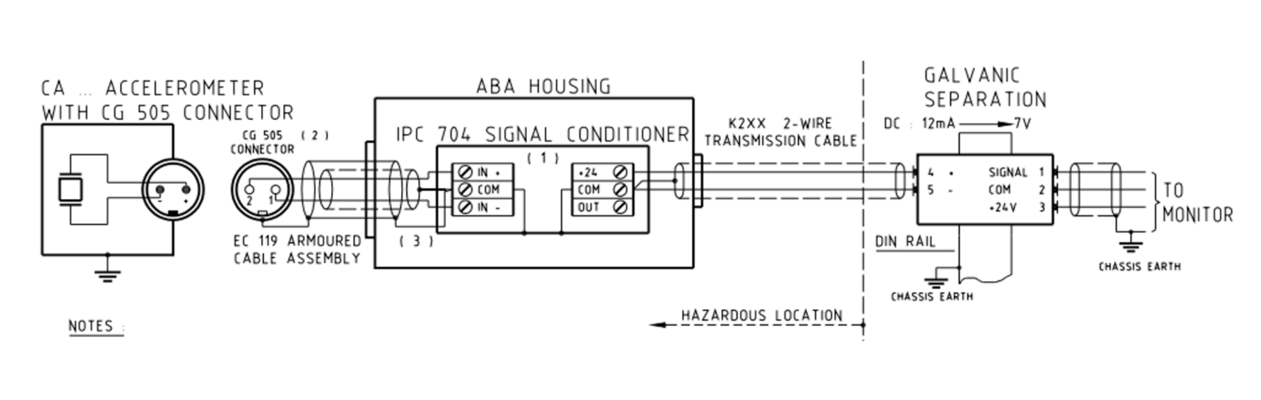

Figure 2: Piezo-electric measurement chain showing the use of low-noise cable (EC119) between the accelerometer and signal conditioner (max. length < 20 m.), and long-distance transmission cable (K2xx) between the signal conditioner and the galvanic separation unit. This arrangement can be used to support long cable lengths, hazardous area installations, or both.

It is also worth noting that although cable specially engineered to better withstand elevated radiation environments is available, its distributed capacitance is usually higher than cables designed for long transmission distances. vibro-meter's K219 cable with Radox® insulation is one example of such cable. If frequency response is of concern, this cable would generally be used only to carry the signal to the boundary of the containment area, and then transitioned to other cable for the remaining distance if quite long.

Cable routing is important for signal quality. Whenever possible, measurement cables should not go along the power cables in a power station, they should be segregated in separate cable trays. The high currents in power cables may induce noise into the "low-voltage" cables, even with shield properly grounded, depending on the cover performance of this shield. If such proximity cannot be avoided, then crossing perpendicularly to the power cable is preferable than going along parallel to it.

1.2 Measuring in water

Water has a major effect on the so-called "apparent capacitance" of cables. When a cable is submerged in water, it will shrink slightly. Although this slightly reduces the distance between wires, and thus the capacitance, this is not the primary contributor to the overall capacitance change. Instead, it is the contribution of the water itself that now surrounds the overall cable. This is not to say that the conductors themselves are now wet and that water has permeated the inside of the cable. The cable remains dry but what surrounds its outermost jacket is no longer air – it is water – and this changes its apparent capacitance. This occurs because the electric field surrounding the conductors does not constrain itself simply to the inside of the overall cable assembly; it extends to the environment surrounding the cable. If that environment changes, so do the properties of the electric field.

Consider that capacitance depends mostly on the relative permittivity[2] (ε_r) of its dielectric material. For air and most plastic materials used in cable insulation, the relative permittivity (ε_r) is approximately 1.0 (to within 3 significant decimal places). In contrast, the relative permittivity of water is nearly eighty times larger, or $\varepsilon_r \approx 80$. Even though the permittivity of the cable insulation and the interior of the cable assembly remains unchanged, through environment outside the cable has thus changed by a factor of 80. It should thus not be surprising that the apparent capacitance seen by the wires inside is different. Although the effect is complex to precisely quantify by calculation, and will not be attempted here, it can easily be observed in practice: as a length of cable

[2] relative to the permittivity of the vacuum ε_0.

is progressively submerged in water, perhaps ten meters at a time, the attenuation of high frequencies becomes noticeably more and more pronounced. Fully submerged pumps (where sensors and wiring are underwater and must run appreciable distances) typically suffer from frequency attenuation due to this issue. One of the remedies is to use a BOA protection, as provided as an option to vibro-meter cables, on relatively short distances, in order to protect from the proximity of water, which minimises the capacitance. A more efficient remedy is to employ a different signal transmission scheme where a given amount of capacitance has less impact on the signal. This is discussed at greater length in §3.

1.3 Ground loops

Another issue of concern with long transmission distances separating the sensor and the monitoring system is that of differing ground potentials. With such long distances – hundreds of meters or even several kilometers – the ground references used at opposite ends of the transmission line will be somewhat different, and this small voltage difference will cause a current to flow, superimposed on the intended signal. Practically, this will most often be noticeable at the main power supply frequency: 50 Hz or 60 Hz, along with possible harmonics of it.

If the sensor is totally isolated from the ground, and the entire measurement chain plus field wiring tied only to ground at the monitoring system, one might be prone to assume that ground loops will not occur because it is tied to a "clean" ground at only one end – not both. However, problems can still occur due to capacitive coupling between the sensor and the machine to which it is attached. The sensor is thus not truly isolated, and the measurement chain will exhibit the effects of a so-called "dirty" ground at the sensor end of the chain and a different "clean" ground at the monitoring system. This often occurs with proximity chains and IEPE-type accelerometer and piezo-velocity "case isolated" sensor chains.

For non-isolated sensors, the sensor has an electrical conduction path to its mounting surface (usually the machine) and the shield will have continuity with this local ground. If the corresponding monitoring system has a purely differential input (like the inputs of the MPC4^{Mk2} module in our VM600 platform), this will not pose a problem: any signal common to both the signal and ground terminals will be rejected in the monitor's first stage of input electronics and only the differential signal between the two will be passed. The key parameter for quantifying this rejection is the "Common Mode Rejection Ratio" (CMRR). For example, the MPC4^{Mk2} has a CMRR of > 60 dB, which is sufficient in most cases.

In the rare instance that this CMRR is insufficient, or if the ground difference goes beyond the common mode acceptance (±50 Volts for the MPC4^{Mk2}), a galvanic separation device can be used, such as vibro-meter's GSI127, which increases both the CMRR and acceptable common mode input voltage range while providing a 0-20kHz frequency bandwidth.

Regardless, the presence or absence of a ground loop should be assessed as part of system commissioning at site and mitigated if present. Ground loops are highly installation-dependant and cannot always be anticipated. In addition, the remedy for a ground loop caused by common mode effects is the opposite of the remedy for that caused by capacitive coupling effects, and when a ground loop represents contributions of both effects, a compromise may be necessary. This is one reason why vibro-meter's accelerometer cables have a provision for either connecting or isolating the shield from ground at the sensor location. In some instances, the shield will perform best when disconnected from ground; in others, it will perform best when connected to ground. With this flexibility, various situations can be accommodated with a single cable. vibro-meter's EC-318 cable used with IEPE-type accelerometers is one such example.

2

LONG TRANSMISSION OF DYNAMIC SIGNALS IN VOLTAGE MODE

The voltage transmission mode is well known and widely used for transmission of vibration signals from proximity probes, accelerometers, velocity sensors, pressure sensors, speed sensors, LVDTs, and other machinery monitoring transducers. Here, we examine the limitations of this technique in terms of frequency response and its corresponding limitations to provide reliable signals over long distances.

2.1 Equivalent circuit diagram

Figure 3 shows the equivalent circuit diagram for modelling transmission cable.

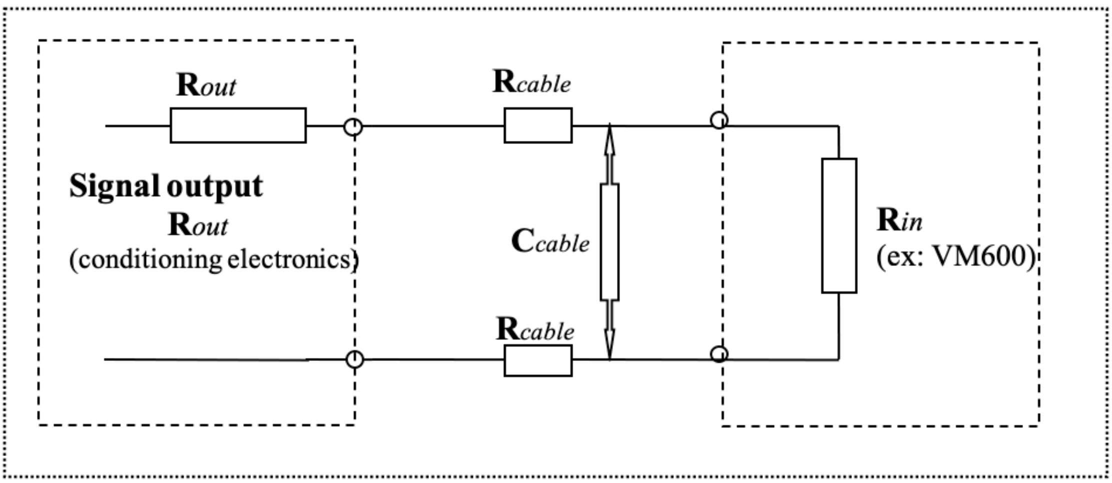

Figure 3: Equivalent circuit diagram for a transmission cable carrying a dynamic signal. There may or may not be another wire for carrying sensor power. Also, the conditioning electronics may be integral with the sensor or external to it. For simplicity, the cable shield is not shown and only the signal conductors are depicted.

The complete impedance calculation is complex, but in practical terms, we have two simple effects:

1. the cable resistance forms a voltage divider with the other resistances, which acts to decrease the sensitivity;
2. the cable distributed resistance and distributed capacitance act as a low-pass filter, which leads to an attenuation of the high-frequency content[3].

[3] When computing the frequency response limitations incurred by transmission cable, the distributed inductance of the conductors as well as input/output devices are rarely large enough to be of concern and can be safely omitted from the calculations. However, this is not the case for intrinsically safe installations in hazardous areas where maximum values for inductance and capacitance are of concern. The inductances of the field devices and the transmission cable must be quantified to ensure they remain within the allowable limits imposed by the relevant certification agency. For additional information, see also References [3] and [4].

2.2 Reduction of sensitivity

The voltage divider effect is computed using Ohm's Law and affects both the AC and DC parts of the signal. If no cable resistance was present, the voltage divider would be simply:

1

$$V_{in} = V_{out}\left(\frac{R_{in}}{R_{in} + R_{out}}\right)$$

Where V_{in} is the voltage "seen" by the monitor and thus the "apparent" sensitivity of the sensor. However, since cable resistance is not zero, our voltage divider equation becomes:

2

$$V_{in} = V_{out}\left(\frac{R_{in}}{R_{in} + 2\,(R_{cable}) + R_{out}}\right)$$

It can be seen that in an ideal circuit where both R_{out} and $R_{cable} = 0$, equation (2) reduces to $V_{out} = V_{in}$ and all of the voltage produced at the sensing apparatus is seen by the monitor. However, for real sensing apparatus and real transmission cables these values are not zero and the error they introduce becomes the resistances they present in the circuit as a percentage of the total resistance:

3

$$Error = \frac{R_{out} + 2\,(R_{cable})}{R_{in} + 2\,(R_{cable}) + R_{out}}$$

The resistance of the cable now adds to the output resistance of the conditioner, both ways (hence, R_{cable} appears twice in the equation).

Depending on the input impedance of the monitoring module, the resistance $\boldsymbol{R}_{in}$ will differ. Although one approach is to design the input with a very high impedance (resistance) so the effects of R_{out} and R_{cable} are minimised, this has the undesirable effect of decreasing the input signal power where:

4

$$P_{in} = \frac{(V_{in})^2}{R_{in}}$$

As can be seen, an increase in the input resistance R_{in} results in a directly proportional decrease input signal power. For example, a doubling of R_{in} results in a halving of P_{in}. This weaker signal power means that the circuit is now more vulnerable to noise, whether the noise intrudes via the cable or elsewhere, and the circuit's overall signal-to-noise ratio is degraded.

For reliable transmission, the input impedance thus cannot be too high. A value between 100 kΩ and 1 MΩ is generally advised.

From the other side, the output impedance (here $\boldsymbol{R}_{in}$) cannot be too small either. A certain resistance is required for the protection of the electronics, and for the stability, in particular in high frequencies, where oscillations can take place.

A quick calculator is available for this Ohmic loss on the internet in [6].

Practical example 1:

Task:
Determine the error introduced by 30 m of K309 cable when using a vibro-meter IPC707 signal conditioner into a VM600 MPC4^{Mk2} module using voltage-mode transmission.

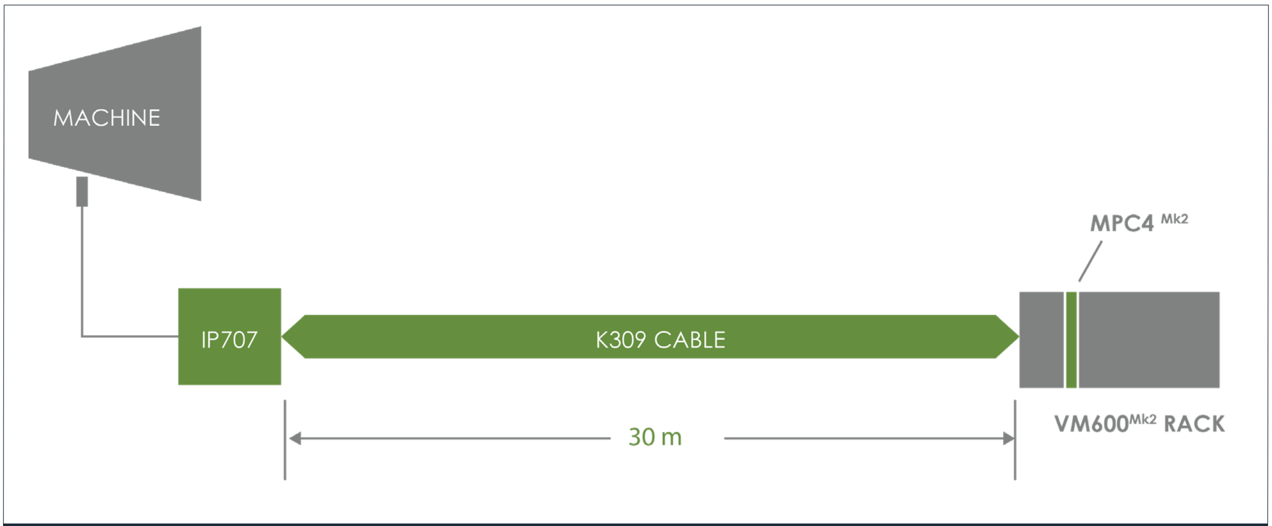

Figure 4: Arrangement for Practical example 1 – part 1.

Repeat for 5 km of K309 cable.

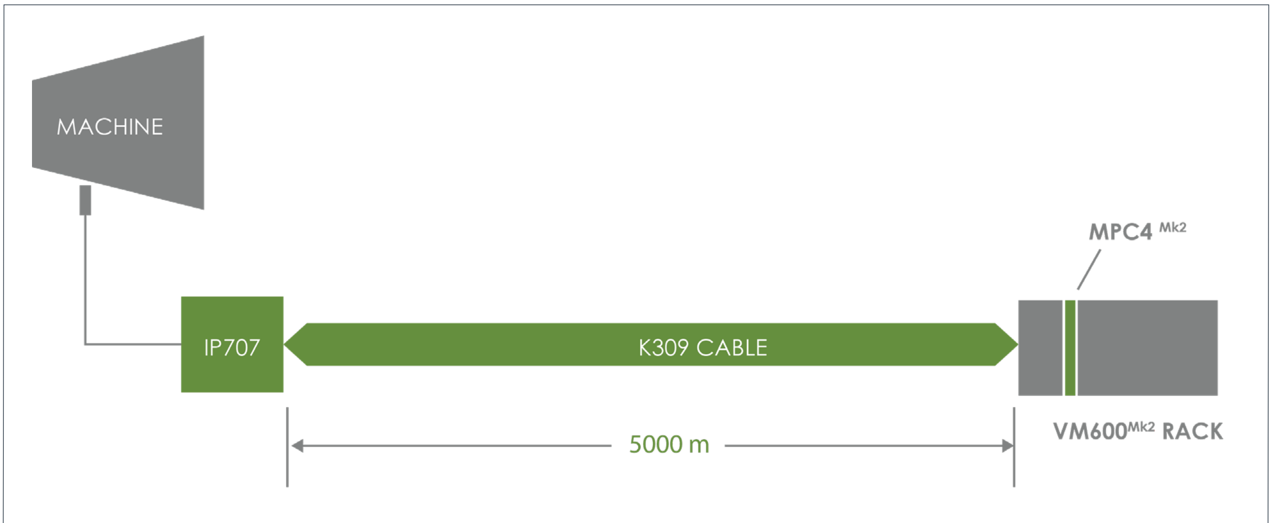

Figure 5: Arrangement for Practical example 1 – part 2.

Steps:
Referring to the relevant datasheets, we obtain the following values:

R_{out}	<	500 Ω	(IPC707)
R_{cable}	=	L x 56 Ω/km	(K309)
R_{in}	>	100 kΩ	(MPC4^{Mk2})

For 30 m (0.03 km) of cable, the error using equation 4 is:

$$Error = \frac{R_{out} + 2\,(R_{cable})}{R_{in} + 2\,(R_{cable}) + R_{out}} = \frac{500 + 2\,(56 \cdot 30/1000)}{100{,}000 + 2\left(56 \cdot \frac{30}{1000}\right) + 500} = 0.005 = 0.5\%$$

With 5 km of cable the error becomes:

$$Error = \frac{R_{out} + 2\,(R_{cable})}{R_{in} + 2\,(R_{cable}) + R_{out}} = \frac{500 + 2\,(56 \cdot 5000/1000)}{100{,}000 + 2\left(56 \cdot \frac{5000}{1000}\right) + 500} = 0.0105 = 1.05\%$$

Even though 5km of cable introduces an error in excess of 1%, and thus outside of published system accuracy specifications, a fully digital system such as the VM600 will generally allow the user to configure the input sensitivity and compensate for this effect, both on the AC and DC components of the signal.

Practical example 2:

Task: *Describe the effect of a Zener barrier in the circuit of Figure 4.*

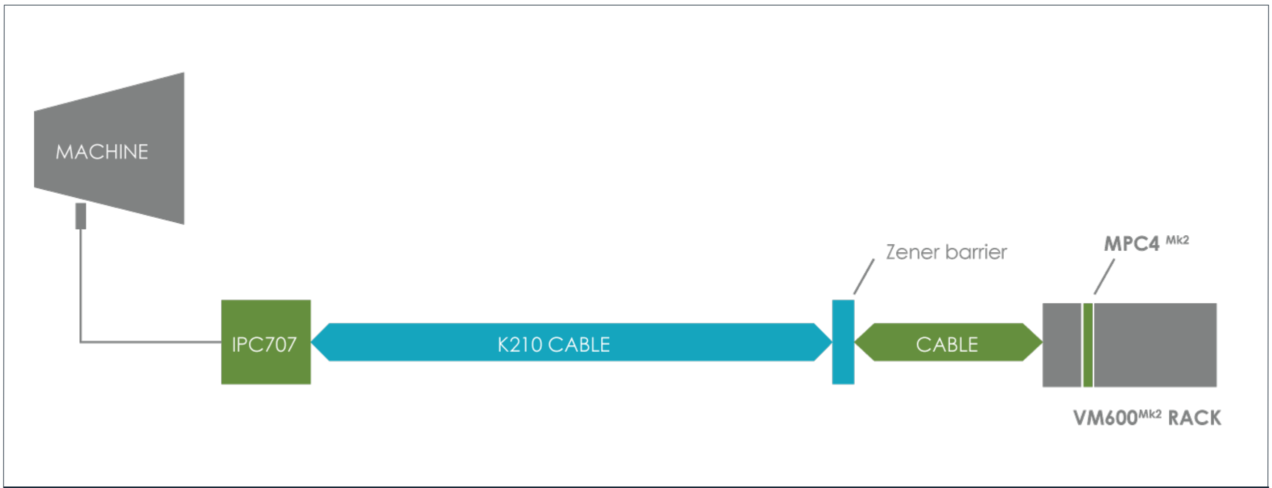

Figure 6: Arrangement for Practical example 2.

Steps:
When a Zener barrier is used to achieve intrinsic safety for a measurement chain in a hazardous area, the barrier also represents a resistance ($R_{barrier}$) and corresponding voltage drop, thus introducing another degradation of the output sensitivity from the conditioning electronics[4] as shown in Figure 4. The equation is modified by simply adding $R_{barrier}$ to the denominator as below.

5

$$V_{in} = V_{out}\left(\frac{R_{in}}{R_{barrier} + R_{in} + 2\,(R_{cable}) + R_{out}}\right)$$

[4] Compensating for the voltage drop across a Zener barrier is another use of the configurable sensitivity in the monitoring module, allowing the user to mitigate these sources of error and achieve better system accuracy.

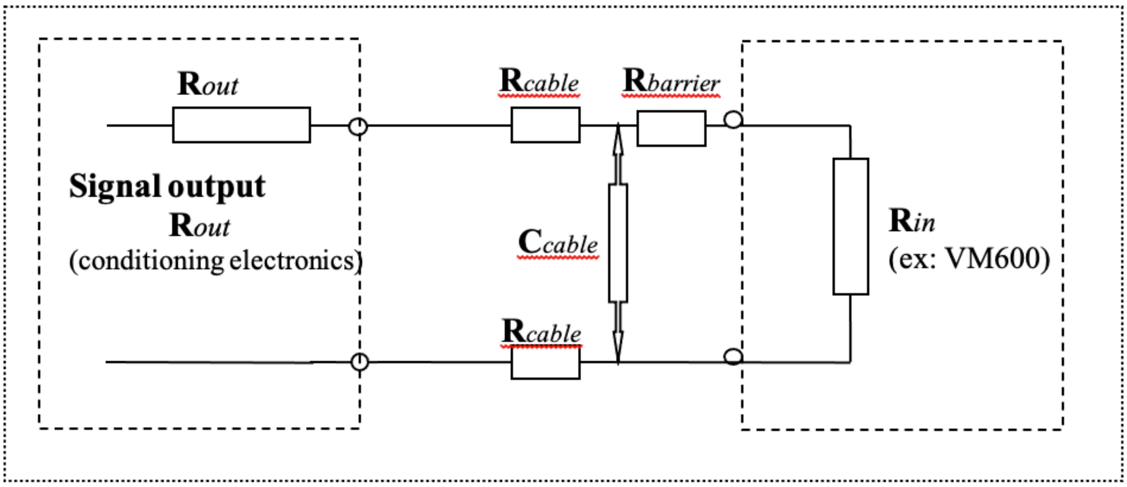

Figure 7: The revised equivalent circuit diagram when a Zener barrier is added to the arrangement of Figure 3.

The resistance of the barrier, then adds to the error, aside the resistance of the cable.

2.3 Power supply check

For all transmission systems, the Ohmic loss shall also be evaluated on power supply voltage as well as on the signal. This time, it shall be checked that the voltage remaining at the device is enough to power it. This applies to voltage transmission as well as current transmission modes.

Let us assume that the power supply is given as:

PS in Volt_DC

A certain part is lost in the cable as:

$-R_{cable} \times I$ for each wire.

We shall take the worst case: corresponding to the maximum current as: $I = I_{max}$.

Then the remaining voltage is:

$$PR = PS - 2 \times R_{cable} \times I_{max}$$

And shall be larger than the necessary power for the device, so:

$$PS - 2 \times R_{cable} \times I_{max} > PS_{dev}$$

Hence the limitation on the cable resistance:

$$Rcable < \frac{PS\text{-}PSdev}{2\ .\ Imax}$$

Example:
- For a proximity chain (Voltage mode), composed of TQ902 probe and IQS9xx conditioner
- The IQS9xx requires -19 Volts minimum (as documented in the datasheet for Voltage mode).
- The power supply provided by VM600[Mk2] is -24 Volts
- The IQS consumes up to 25 mA.

Which length of cable K309 is acceptable?
The resistivity of the cable is 56 Ω/km.

- From the equation above, we have the limit:
  - $R_{cable} < (24 - 19) / (2 \times I_{max})$

Which gives:
  - $R_{cable}limit = 100\ \Omega$

- So the limitation on the length is:
  - $L^{limit} = 100 / 0.056 = 1780$ *meters*

This is quite acceptable in comparison to other limitations on the frequency band.

2.4 Attenuation of high frequencies

The cable's internal distributed capacitance (C_{cable}) combined with the sensor chain's output resistance (R_{out}), and to a lesser extent its internal resistance (R_{cable}) creates a low-pass filter. Because a low-pass filter, by definition, does not affect low-frequency signal components, the DC part of the signal is not affected; the filter affects only the AC signal component. Further, only the AC component around and above the so-called cut-off frequency (F_c) is attenuated, where F_c is given by[5]:

6

$$F_c = \frac{1}{2\pi\,(R_{out} + 2\,R_{cable})\,.C_{cable}}$$

The cut-off frequency corresponds to an attenuation of -3dB for a first-order filter, and this corresponds to roughly -30%. Modelling the circuit of Figure 3 as a first-order low-pass filter gives a "worst case" (i.e., most conservative) result with classical "roll off" (attenuation) of 20 dB/decade (equivalent to 6 dB/octave). For a more accurate calculation, the circuit can be modelled by a continuous equation. For simplicity, that equation is not shown. Be aware that the first-order model is a simplification and that the real circuit will generally behave differently than this approximation of the first-order model.[6] Notably, the attenuation by a real cable starts before the so-called cut-off frequency, but there the slope is not as steep as the slope of the first-order filter, it is smoother. Beyond the cut-off frequency, this is the opposite: the slope will accelerate and be steeper than the equivalent of a first-order filter: more than 6 dB/octave. For the user then, depending on the accuracy required, the first order approximation will underestimate the attenuation in the narrow zone: if for example a few percent is specified. Similarly, at much higher frequencies than the cut-off, also the cable with distributed capacitance and resistance, will behave more like a second order filter or more. However, the limited zone around the cut-off frequency is well modelled as a first-order filter, and the curve departs from this model way below and way above this frequency.

[5] This formula is only used for voltage-mode transmission.
[6] Those unfamiliar with the basics of a low-pass filter, such as roll-off, passband, stopband, and cut-off frequency are encouraged to consult reference [5].

In other sources, only 1.R instead of 2.R is considered in this calculation. 2.R corresponds to the way in and way out of the voltage from the source. Omitting one way underestimates the cut-off-frequency, but this is over an over-estimated value already. As we want to be conservative, and insure a correct accuracy of the signal transmission, we will consider 2.R in this calculation.

First of all, the determination of the "effective" distributed capacitance can be estimated by various methods.

- Ideally the capacitance between 1 core and all the remaining conductors, clamped together is measured on the cable and given. The conductors clamped together are: other cores and the shield (one core for a pair and 2 cores for a triad). In the example, for vibro-meter cables K210 and K310, this data is provided in the datasheet. Then this value is to consider in the following calculation,

- else, if this information is not given, it can be estimated by the sum of:

  - the capacitance between a conductor core and the overall shield (Cc-s) "core-to-shield"
  - and the core-to-core capacitance ($C_{c\text{-}c}$) between two active signal conductors,
  - for a triad: the capacitance ($C_{c\text{-}c}$) shall be counted twice, because there are 2 other core conductors in the cable.

We have:

7 For a triad:

$$\boldsymbol{C_{cable} \cong 2.C_{c-c} + C_{c-s}}$$

And:

13 For a pair:

$$\boldsymbol{C_{cable} \cong C_{c-c} + C_{c-s}}$$

In some cases of cables datasheets, ($C_{c\text{-}s}$) might be missing, and replaced by the value of capacitance of: all cores clamped together against the shield ("Cores-to-shield"). The estimation can still be done with these figures, but involves more calculations.

Eventually, the effective total capacitance of the cable is obtained by multiplying the effective distributed capacitance by the length of the cable.

Practical example 3:

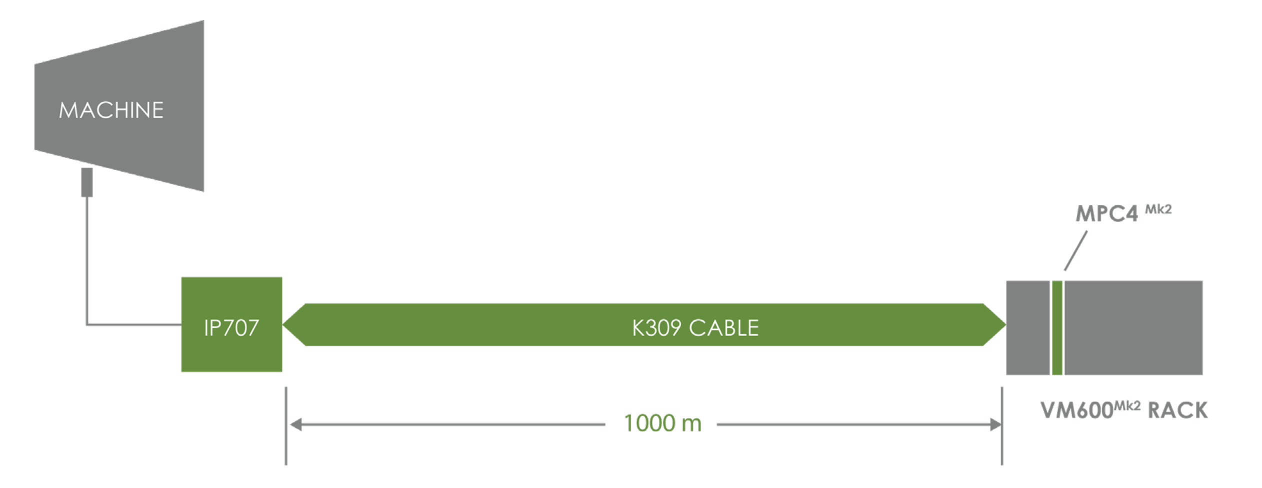

Figure 8: Arrangement for Practical example 3.

Task: *Using the same cable type (K309) and devices as in example 1, calculate the cut-off frequency for 1 km of cable.*

Steps:
Determination of the "effective" capacitance involves both the core-to-core capacitance ($C_{c\text{-}c}$) between active signal conductors, and between all the conductor cores and the overall shield ($C_{c\text{-}s}$). From the K309 datasheet:

Between wires:	$C_{c\text{-}c}$ = 150 pF/m
Between all wires together and shield:	$C_{c\text{-}s}$ = 270 pF/m

For a triad cable, these values combine using the following approximation that accounts for the increase in capacitance due to capacitive interaction between the conductors and the shield:

7

$$\boldsymbol{C_{cable} \cong 2.C_{c-c} + C_{c-s}}$$

In our example:

$$C_{cable} \cong 2.C_{c-c} + C_{c-s} = 2 * 150 + 270 = 570\,pF/m$$

For 1000 m of cable, our values become:

C_{cable} = 570 *p*F/m x 1000 m = 570 *n*F = 0.57 *μ*F
R_{cable} = 56 Ω/km x 1 km = 56 Ω
R_{out} = 500 Ω

We then use equation 6:

$$F_c = \frac{1}{2\pi\,(R_{out} + 2\,R_{cable})\,C_{cable}} = \frac{1}{2\pi\,(500 + 112)\,(0.57\;x\;10^{-6})} = 456\ \text{Hz}$$

Unfortunately, this will almost always be insufficient for vibration protection and monitoring except for shaft-relative vibration on very low-speed machines (such as hydro units) or for thrust position measurements (~DC). In addition, this frequency band does not comply with ISO 20816-x standards for protection where a minimum passband of [10..1000 Hz] is required. A better cable can probably be found, but for a higher cost of installation.

Please note that, considering relative vibrations, vibro-meter's newest signal conditioner for proximity probes – the IQS900 – could be used in this situation with acceptable results for shaft-relative vibration. This is because R_{out} is only 100 Ω for the IQS900 instead of 500 Ω for the IPC707. Solving for F_c using this revised value for R_{out} yields a cut-off frequency of 1320 Hz, which satisfies ISO requirements. Thus, up to 1 km of this cable can be supported for voltage-mode shaft-relative measurements when using the IQS900 and K309 transmission cable.

2.5 IEPE mode limitation (constant current power supply)

For the IEPE[7] mode of transmission, which is very popular for accelerometers, these effects are not the only issues to consider.

The IEPE technique is a 2-wire transmission scheme in which the sensor power and the sensor signal share the same loop (i.e., conductors pair). A constant current energizes the sensor and the sensor's dynamic signal is returned as a dynamic voltage.

In §2.3, we examined the low-pass filtering effect of field wiring via a simplified first-order model. This effect exists also in the IEPE mode of transmission, but is less pronounced than in pure 3-wire voltage-mode circuits.

When IEPE accelerometers[8] are used with excessively long field wiring, signal asymmetry can occur as shown in Figure 5.

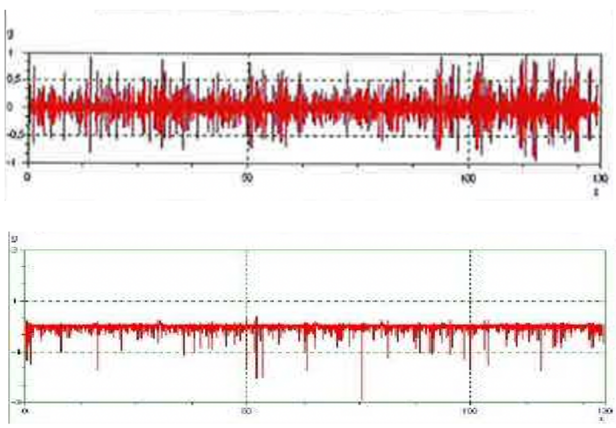

Figure 9: The top trace shows the signal measured at the IEPE sensor's output, before field wiring is attached. The bottom trace shows the effect of field wiring, resulting in signal asymmetry. This asymmetry occurs when the sensor's energizing current dynamically drops below its minimum threshold due to cable capacitance.

[7] Integrated Electronic Piezo Electric. Also known as CCLD (Constant-Current Line Drive) as well as various trademarks such as ICP®, Isotron®, and DeltaTron® where the trademarks are property of their respective owners.
[8] These effects are also present with so-called "Integrating IEPE Accelerometers" providing a velocity output.

In §2.3, we saw that the cable capacitance was detrimental to the frequency response and that larger values of capacitance had the effect of lowering the cut-off frequency F_C. Here, the same cable capacitance also plays a destructive role in the signal integrity, but via a different mechanism.

The IEPE sensor needs a certain amount of energizing current to operate normally (4 mA is typical). When insufficient energizing current is delivered – even instantaneously - the sensor's output will be incorrect. Depending on the amplitude of the dynamic signal, a fraction of this energizing current will be stored in the capacitance along the cable. For proper operation, there must always be a margin between the power delivered and the power used by the sensor. For example, if the system (such as the MPC4^{Mk2} module in our VM600) delivers a nominal constant current in excess of 6 mA, and the IEPE accelerometer requires a minimum of 4 mA for proper operation, the margin is 2 mA. This is the current that can be "wasted" in the cable.

The temporary storage of energy by this cable capacitance is proportional to the voltage difference; this is why the effect depends on the signal itself and is therefore non-linear. The effect can be clearly seen in Figure 5 where the positive part of the signal is affected but the negative part is not. This is because sufficient energizing current is only "stolen" by the cable capacitance during the positive part of the waveform.

Calculation of this effect involves understanding the sensor's voltage output (V), which in turn depends on the measured acceleration[9].

Consider that the storage of electric charge (Q) is:

$$Q = C_{cable} \times V \quad (8)$$

The current (I^A) is the rate of change (i.e., derivative) of this quantity of charge:

$$I_A = \frac{dQ}{dt} = C_{cable} \times \frac{dV}{dt} = C_{cable} \times sensitivity \times \frac{dA}{dt} \quad (9)$$

Where *A* is the dynamic acceleration acting on the sensor and sensitivity is the sensor's characteristic relationship between its acceleration and its output voltage (typically 100mV/g).

Note that the static voltage (i.e., bias voltage) does not matter in this calculation because it is a constant and the rate of change (derivative) is thus 0. Once the sensor connection is made to its power source, the bias voltage stabilizes almost instantaneously and from that point forward, only the rate of change of V will affect the current.

The net effect is that a threshold will exist on the maximum rate of change in voltage – the so-called a slew-rate effect. As soon as a sufficiently fast variation of g (and thus V) acts on the accelerometer, it will not be able to deliver a sufficiently fast change in voltage. The signal will be distorted on the positive side. However, the same can not be said of fast deceleration (negative slew rate) because – unlike acceleration – does not act to subtract from the necessary energizing current. As such, the response is non-symmetric and the corresponding effect on the signal's frequency content (i.e. spectrum) cannot be predicted. However, one can still have a rough idea about the frequency above which this effect will start to become apparent.

[9] IEPE devices are not limited to measurement of acceleration and can include force, pressure, and velocity (integrated acceleration) transducers. For purposes of this discussion, however, we have assumed that the IEPE device is an accelerometer.

We do this by considering the maximum acceleration amplitude (A_f) that can be tolerated at a particular frequency (f) and by assuming a sinusoidal variation of acceleration as follows:

$$\frac{dA}{dt} = A_f \, x \, 2\pi f \tag{10}$$

and:

$$\frac{dA}{dt} = \frac{I_A}{sensitivity \times C_{cable}} \tag{11}$$

Using 10 and 11 to solve for f, we obtain:

$$f = \frac{I_A}{2\pi \times A_f \times sensitivity \times C_{cable}} \tag{12}$$

It can thus be seen that this frequency is inversely proportional to both the capacitance of the cable and to the maximum amplitude expected.

Practical example 4:

Task: *Calculate the approximate frequency at which signal degradation will begin to occur from a 100 mV/g IEPE accelerometer requiring a minimum constant current of 4mA and using 1 km of K209 cable if a maximum of 10 g's are expected. Assume that it is connected to an MPC4^{Mk2} card in a VM600 rack and that ISO 10816-n passband requirements must be met: up to 1000 Hertz.*

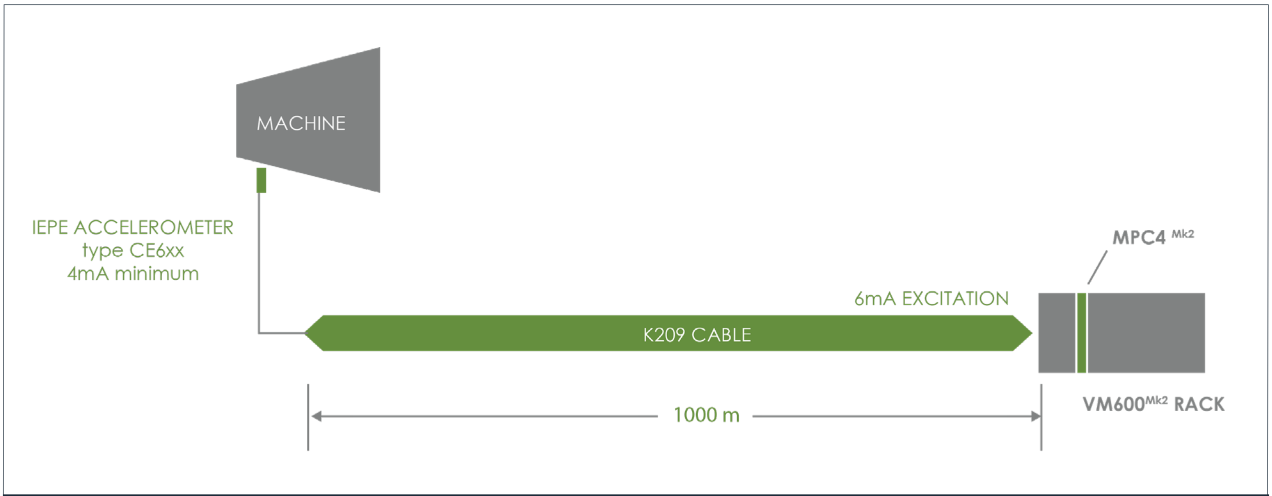

Figure 10: Arrangement for Practical example 4.

Steps:
Consulting the K209 datasheet, we obtain the following values:

Between wires:	$C_{c\text{-}c}$ = 150 pF/m
Between all wires and shield:	$C_{c\text{-}s}$ = 270 pF/m

Because we are now using a 2-conductor cable instead of a 3-conductor cable, we no longer use equation 7. Instead, we use the equation below:

$$C_{cable} \cong C_{c-c} + C_{c-s} \tag{13}$$

For 1000 m of K209 cable, equation 13 yields a total capacitance of 0.42 μF as follows:

$$C_{cable} \cong C_{c-c} + C_{c-s} = 150\,p\mathrm{F/m} + 270\,p\mathrm{F/m} = 420\,p\mathrm{F/m}$$

$$420\,p\mathrm{F} \times 1000\,\mathrm{m} = 420\,n\mathrm{F} = 0.42\,\mu\mathrm{F}$$

Recalling that the MPC4$^{\mathrm{Mk2}}$ supplies 6mA and that the accelerometer requires a minimum of 4mA, this leaves 2mA of "headroom" that can be sacrificed to cable capacitance. We then use equation 12 to find:

$$f = \frac{I_A}{2\pi \times A_f \times sensitivity \times C_{cable}} = \frac{0.002\ \mathrm{A}}{2\pi \times 10\ \mathrm{g\ x}\ 0.1\frac{\mathrm{V}}{\mathrm{g}} \times 0.42 \times 10^{-6}\ F} = 758\ \mathrm{Hz}$$

This is not sufficient to meet the requirements of ISO 10816-n ([10..1000] Hertz), unfortunately.

While a better quality of cable could be chosen (such as the vibro-meter K210, more costly), another solution could be to locate the power supply (constant current mode) closer to the sensor and thus shorten the required field wiring length. However, one would also have to consider the low-pass filtering effects discussed in §2.3 and perform the necessary calculations using the appropriate resistance and capacitance values for K209 cable along with R_{out} for the MPC4$^{\mathrm{Mk2}}$. This will yield a maximum allowable cable length to avoid low-pass filtering effects. The shorter of the two length constraints would then be selected to meet ISO 10816-n: one based on the slew-rate limitations calculated above and one based on frequency response limitations (§2.3).

3

LONG TRANSMISSION OF DYNAMIC SIGNALS IN CURRENT MODE

This method uses a varying current instead of varying voltage to represent the dynamic signal. In theory, this mode eliminates the effect of Ohmic losses from long runs of field wiring. While there is still energy loss in the line, it affects only the voltage at the destination – not the current.

In current-mode signal transmission, the capacitance of the cable still matters, but far less than in voltage-mode. However, even in a current-mode system, the receiving device (the monitor) reads the current on a so-called "reading resistor" as a voltage (V = I x R), so there still is a voltage involved. We examine this influence in the examples that follow, using two typical vibro-meter measurement chains.

3.1 Example of vibro-meter IPC707 or CE-type accelerometers

The IPC707 is a flexible signal conditioner (so-called "charge amplifier") designed for use with vibro-meter sensors that do not have integral signal conditioning electronics, such as our CA accelerometers and our CP dynamic pressure sensors. It can be configured for voltage-mode or current-mode output. Here, we consider its capabilities when providing a current-mode output.

Our CE accelerometers have integrated signal conditioning and thus differ from our CA models. Consequently, they do not require the use of a device like the IPC707: the typical output of 5 to 50 μA per g can be received directly by the monitor. Many of our CE models use a special current transmission technique, overcoming some of the limitations of IEPE devices which are inherently voltage-mode devices and hence vulnerable to the issues discussed in § 2.3 and §2.4. This technique offers several benefits, including:

- the power supply issue of §2.4 does not exist because the power supply is regulated in the remote electronics and the power margin is always ensured, regardless of signal amplitude or frequency;
- the frequency response is increased for an equivalent length of transmission cable, or, similarly, possible length of transmission cable is increased for an equivalent frequency band.

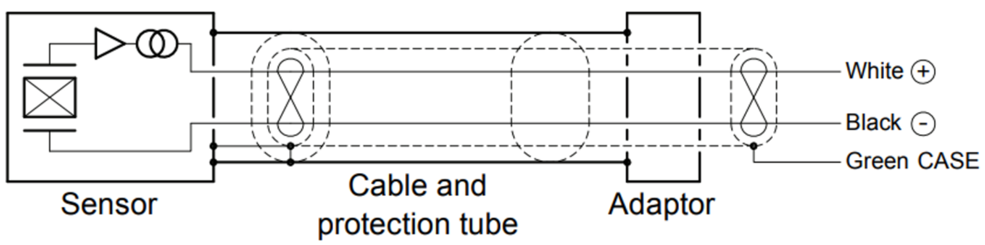

Figure 11: Schematic of a typical vibro-meter CE accelerometer. Note that for improved immunity, the local cable has an inner and an outer shield, joined at the sensor case.

When in current mode, both IPC707 signal conditioners and CE accelerometers show an improvement in capabilities to drive extended lengths of field wiring. Examples of CE accelerometers are CE134, CE311 and CE281 in the present vibro-meter offer in Energy sector.

In current mode, there still is a low-pass filtering effect due to the cable capacitance; however, the effect is smaller than in voltage mode.

Referring to Figure 3, the computation of cut-off frequency is analogous to that used in equation 6, but we concern ourselves with the input impedance (R_{in}) of the monitor rather than the output impedance (R_{out}) of the signal conditioner. This is because we are not concerned with the voltage leaving the signal conditioner – we are concerned with the *current* leaving the power supply from the monitor. We are, however, still concerned with the voltage arriving at the receiving device (monitor) and this is affected by the cable resistance and the resistance of the monitor's reading resistor (R_{in}). Equation 6 (used for voltage-mode signal transmission) is thus transformed as shown below for current-mode signal transmission:

14

$$F_c = \frac{1}{2\pi\,(R_{in} + 2\,R_{cable})\,C_{cable}}$$

The capacitance of the effective cable C_{cable} is calculated in the same way as in §2.3 depending on the information given by the cable manufacturer. The current loop transmission technique also allows to use 2 wires (a shielded pair) instead of 3 wires (a shielded triad), and this offers also an advantage, in practice, in terms of effective distributed capacitance, thus in terms of frequency band.

Practical example 5:

Task: *Calculate the cut-off frequency for 1000 m of K209 cable for a measurement chain using current-mode signal transmission into an MPC4^{Mk2} card as the receiving equipment.*

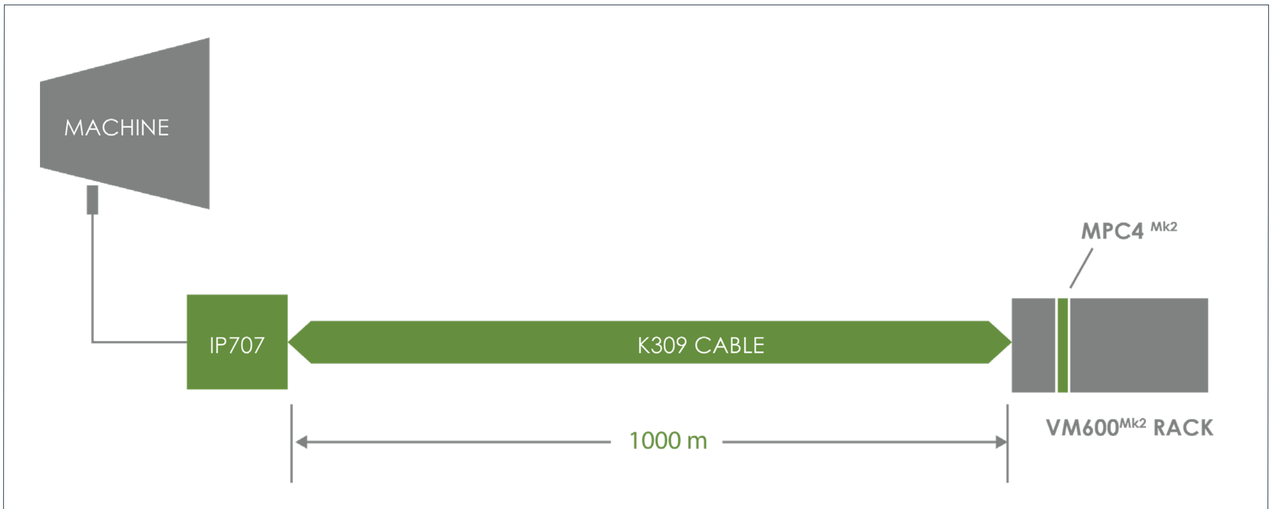

Figure 12: Arrangement for Practical example 5

Steps:
Consulting the datasheet for the MPC4[Mk2], we find that the input impedance for a current-mode input is 200 Ω. From the K209 datasheet, we find that R_{cable} is 56 Ω/km. We then use equation (14) to solve for F_c:

$$F_c = \frac{1}{2\pi\,(R_{in} + 2\,R_{cable})\,C_{cable}} = \frac{1}{2\pi\left(200\,\Omega + 2 \times 1\text{km} \times 56\,\frac{\Omega}{\text{km}}\right) \times 0.42 \times 10^{-6}\ \text{F}} = 1215\ \text{Hz}$$

Using current-mode signal transmission and the same 1 km length of field wiring, we have a significant improvement (+250%) in F_c compared to the voltage-mode signal transmission of example 3. This is now enough to comply with the 1000 Hz requirement of the ISO 10816-x standard for reliable machinery protection.

The improvement is both due to a smaller capacitance of the cable, and a smaller impedance to consider. It is also a significant improvement to the example 4, using an IEPE accelerometer on the same distance.

Practical example 6:

The distance: 1000 feet, is considered as a standard length for O&G applications (API 670).

Task: *Repeat examples 3, 4 and 5, but for 305m (1000 feet) of field wiring[10].*

Steps: Repeating the computations of example 3 with 305 m instead of 1000 m of cable gives:

C_{cable} = 570 pF/m x 305 m x 0.001 nF/pF = 174 nF
R_{cable} = 56 Ω/km x 0.305 km = 17 Ω
R_{out} = 500 Ω

We then use equation 6:

$$F_c = \frac{1}{2\pi\,(R_{out} + 2\,R_{cable})\,C_{cable}} = \frac{1}{2\pi\,(500 + 34)\,(174\,x\,10^{-9})} = 1713\ \text{Hz}$$

Repeating the computations of example 4 with 305 m instead of 1 Km of cable gives:

C_{cable} = 420 pF/m x 305 m x 0.001 *n*F/pF = 128 *n*F
Sensitivity = 100 mV/g
Amplitude A*f* = 10 g

We then use equation (12):

$$f = \frac{I_A}{2\pi \times A_f \times xsensitivity \times C_{cable}} = \frac{0.002\ \text{A}}{2\pi \times 10\ \text{g} \times 0.1\,\frac{\text{V}}{\text{g}} \times 0.128 \times 10^{-6}\ \text{F}} = 2487\ \text{Hz}$$

[10] This length is often considered as the practical upper limit for reliable transmission of vibration signals. However, this assumes voltage-mode rather than current-mode signal transmission.

Repeating the computations of example 5 with 305 m instead of 1000 m of cable gives

C_{cable} = 420 pF/m x 305 m x 0.001 nF/pF = 128 nF
R_{cable} = 56 Ω/km x 0.305 km = 17 Ω
R_{in} = 200 Ω

We then use equation (14):

$$F_c = \frac{1}{2\pi\,(R_{in} +\ 2\,R_{cable})\,C_{cable}} = \frac{1}{2\pi\,(200 +\ 34)\,(128\,x\,10^{-9})} = 5313\text{ Hz}$$

, gives 1700 Hz for voltage-mode transmission (example 3) and 5300 Hz for current-mode transmission (example 5). Thus, for the same hardware, and the same cables (K309 or K209):

- the voltage-mode signal transmission is limited to about 1.7 kHz and further also, in addition, to ~2.5 kHz with IEPE mode,
- the current-mode signal transmission is limited only to about 5.3 kHz.

These examples show frequency response improvement by a factor of more than 3 for current-mode transmission versus voltage-mode transmission. Although the frequency bands are enough for protection in all cases, this difference can be important for machines where gear mesh frequencies, rolling element bearing frequencies, vane passage frequencies, (between 1.7 and 5 kHz) and other signatures are relevant for predictive maintenance and diagnostics.

For the normal cable length (305 m), the current modulation technique permits monitoring and diagnostic of complex machineries, in addition to the protection function, using the same sensor only.

3.2 Solution with vibro-meter GSI technology

vibro-meter's GSI[11] (Galvanic Separation Instrument) device is designed for insertion between the sensing apparatus and the monitoring instrument such as a VM600 rack, VibroSmart distributed modules, or any others monitoring system. It provides numerous capabilities including:

- prevents ground loops (up to +-4000 Volts), with a very high Common Mode Rejection Ratio ("CMRR" is specified)
- uses current modulation mode as a standard,
- passes a regulated power supply to the sensing apparatus
- provides a very low-impedance (<30 Ω) input for the sensor signal
- energy limiting for intrinsically safe installations by serving as a safety barrier.
- 0 - 20kHz frequency bandwidth response (unlike other commonly used galvanic separation devices designed only for quasi-static signals)

Versions exist for accelerometers, for proximity probes, and for IEPE sensors.

The combination of isolation, current modulation, and low impedance input and output, makes the GSI particularly recommended for long distance transmission of dynamic signals, as evidenced in the following examples.

[11] vibro-meter currently offers the GSI 127, providing similar but improved capabilities relative to the legacy GSI 122, 124, and 130 devices.

Practical example 7:

Task: *Repeat example 5, but using a GSI 127_B02 device along with the IPC707. The GSI127 is located close to the monitoring system (typically in the same cabinet), and the IPC707 located close to the machine, if not onto the machine itself. The signal is modulated in current along most part of the travel to destination, which makes it more robust to E.M. disturbances, and less subject to frequency attenuation.*

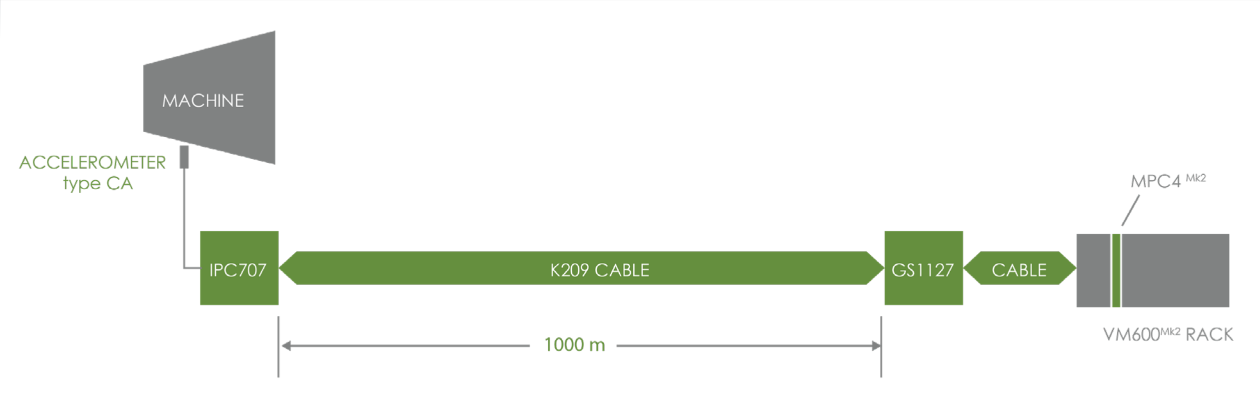

Figure 13: Arrangement for Practical example 7. Note the location of the GSI close to the monitoring rack.

Steps: We use equation 6 as before, but with the input impedance of the GSI 127 (30 Ω) instead of that of the MPC4^{Mk2} (200 Ω). It is considered as R_{out} in the equation, being the output side relative to (or "-seen from-") the front-end.

C_{cable} = 420 pF/m x 1000 m x 0.001 nF/pF = 420 nF
R_{cable} = 56 Ω/km x 1 km = 56 Ω
R_{out} = 25 Ω because 30 Ω is considered only for low frequencies

$$F_c = \frac{1}{2\pi\,(R_{out} + 2\,R_{cable})\,C_{cable}} = \frac{1}{2\pi\,(25 + 112)\,(420\,x\,10^{-9})} = 2766\text{ Hz}$$

Now, by insertion of the GSI 127, we have increased F_c by a further factor of more than 2X (2766 Hz versus 1215 Hz) and this results in more than a factor 6X compared to example 3 in voltage mode; and thereby allowing this measurement chain to be used for monitoring and diagnostics in higher frequencies.

305m (1000 feet) is often taken to be the practical upper limit for field wiring length on voltage-mode transmission. Indeed, if we insert a cut-off frequency of 1000 Hz in equation 6 and then solve for cable length, it can be seen that up to 2350 m of field wiring could actually be supported with K209 cable and the GSI 127_B03 while conforming to ISO 7919.

In addition to cutting the ground loops, and offering an Intrinsic Safety barrier, the GSI improves even further the current modulation technique, increasing frequency content and signal quality over longer distances.

Practical example 8:

Task: *Repeat example 7, but assume 305m of cable. Further assume a 20 kHz pressure transducer is being used for Gas turbine combustion monitoring; and the full bandwidth of the transducer is indeed necessary.*

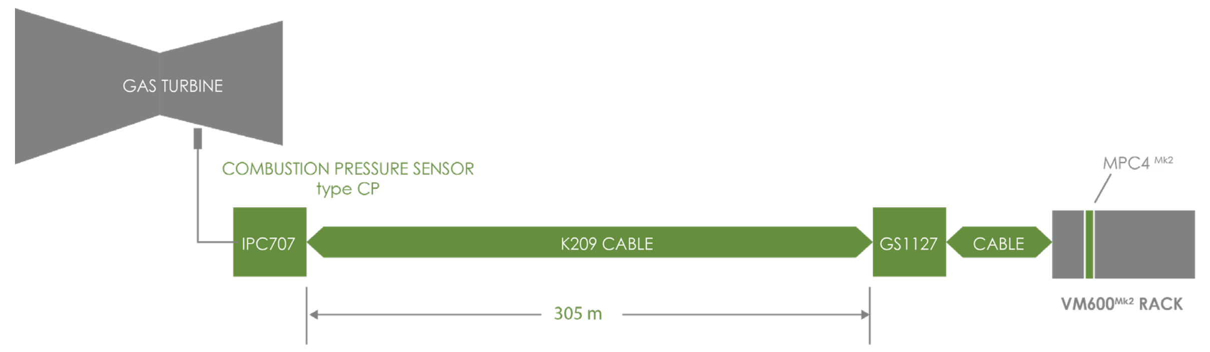

Figure 14: Arrangement for Practical example 8.

Steps: Our values for resistance and capacitance now become:

C_{cable} = 420 pF/m x 305 m x 0.001 nF/pF = 128 nF
R_{cable} = 56 Ω/km x 0.305 km = 17 Ω
R_{out} = 20 Ω

$$F_c = \frac{1}{2\pi\,(R_{out} + 2\,R_{cable})\,C_{cable}} = \frac{1}{2\pi\,(20 + 34)\,(128\,x\,10^{-9})} = 23.2\ \text{kHz}$$

The [0..20 kHz] capability of the sensor, ensuring that the sensor's frequency response is the only thing limiting signal integrity, is then fulfilled by the field wiring. The technique is actually used and opens many possibilities for location of the monitoring system in a convenient and safe location while still allowing high-frequency measurements such as combustion dynamics on gas turbines. Monitoring gearbox casing, rolling element bearings, etc. are also at reach from farther and safer distances in this scenario.

The following plot summarises a computation of frequency versus cable length for 3 different options using actual vibro-meter chains: IPC707, MPC4^{Mk2}, and adding GSI127.

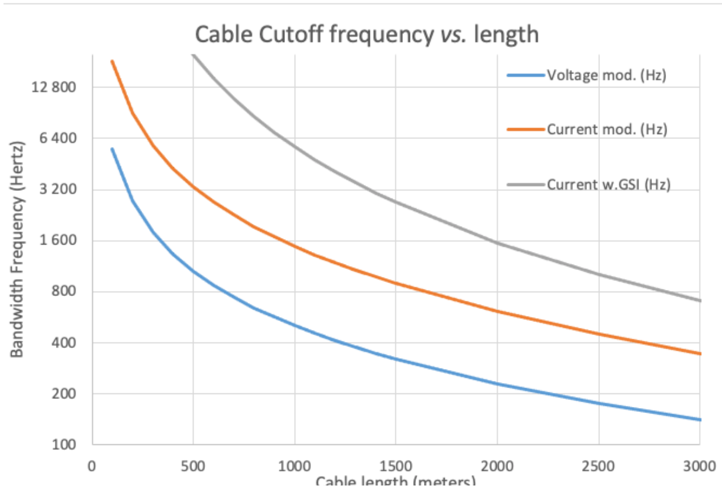

Figure 15: Curves of maximum frequency cut-off for different modulations & solutions: Voltage mode (commonly used), current modulation, and current modulation with the Galvanic Separation Instrument (GSI).

Depending on the example, the benefit of using a current modulation can be roughly a factor 2 to 3 in terms of frequency band; and using the GSI brings another factor 2 to 3 to the bandwidth.

In practice, the problem is to find which frequency band can be accommodated by the needed distance in the plant. The curves network shown above (Figure 7) allows to invert the problem and find which distance permits a given frequency band. Such calculation is given in the downloadable spreadsheet from our website. In this spreadsheet, using MS excel, the data from the components impedances, and the cables characteristics shall be entered, and the curves, as shown above, are displayed. This can be used for accelerometers, proximity probes, speed signals, velocimeters, *etc.*

For example, this calculation sheet can be used to compute the curves for various cables available on the market, and compare their performance depending on their distributed characteristics.

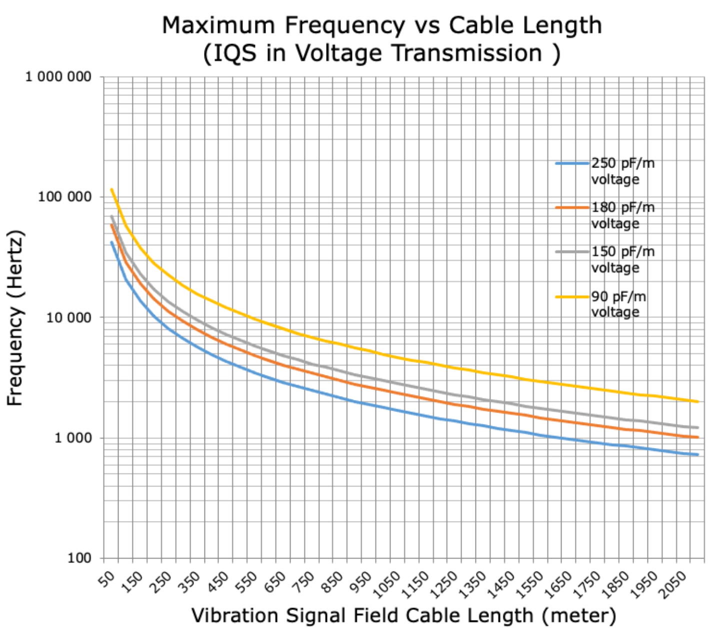

Figure 16: Comparison of cut-off frequency for a vibro-meter IQS proximity measurement chain under voltage-mode transmission, with different cable capacitances.

4

LONG TRANSMISSION USING FREQUENCY MODULATION

4.1 Speed / phase-reference signals

The input to a tachometer can have one or multiple pulses per revolution, obtained from proximity probes, Hall-effect sensors, or magnetic pickups. Phase reference signals assume one pulse per revolution while tachometer signals often use multiple pulses per revolution to improve accuracy as well as reducing the time required to update the readings to a fraction of a revolution.

For proximity probes, the signal amplitude is unaffected by rotational speed and only the period (frequency) of the pulses will change – thus becoming a type of frequency modulated signal where the speed is proportional to the frequency. As with vibration and other dynamic signals, the integrity of speed signals must be preserved and thus the effects of cable length must be examined.

The effects already seen in §2 and §3 apply; namely, reduction of amplitude (voltage-mode transmission only) and low-pass filtering (both voltage- and current-mode transmission) as shown in Figure 7. As was demonstrated, current-mode transmission provides a higher cut-off frequency for a given length of field wiring. Conversely, for a given cut-off frequency, current-mode transmission allows longer field wiring. This has the effect of preserving the speed signal integrity and can be extremely important for proper triggering – particularly in overspeed applications where the consequences of a missed trip can be extremely high.

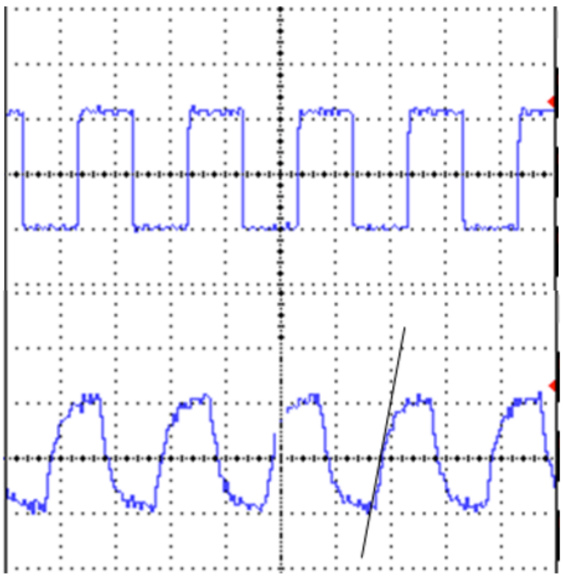

Figure 17: Effect of low-pass filtering (bottom) on a squarewave (top).

The low-pass filtering effect of field wiring acts to smooth the sharp edges[12] of rectangular pulses and give a shape with a slope. This creates two issues with proper measurement:

- It introduces a lag, which may be noticeable depending on where the phase triggering threshold has been set
- It may reduce the amplitudes of pulses, imposing the necessity for a lower threshold level for triggering.

It can be shown that the slope in the bottom trace of Figure 8 is affected by both the size of the shaft discontinuity observed by the probe and by the cut-off frequency imposed by the field wiring.

The response R(*t*) in the time domain is as follows:

15

$$R(t) = D * (1 - e^{-t/\tau})$$

Where D is the depth[13] and decrement "τ" is approximated as:

16

$$\tau = \frac{1}{2\pi . f_c}$$

Equations 15 and 16 can be combined to more clearly see the effect of changes in cut-off frequency:

17

$$R(t) = D * \left(1 - e^{-t/\left(\frac{1}{2\pi f_c}\right)}\right) = D * \left(1 - e^{-(2\pi f_c . t)}\right)$$

With an automatic threshold algorithm or circuitry (such as in the VM600 cards), the frequency can still be measured even though the field wiring attenuates the higher frequencies and thus distorts the crisp edges of the pulses. In other words, the ability to accurately count pulses can be accomplished well beyond what might be assumed based purely on the frequency limit given by the previous calculations in §2 and §3. This approach, however, does have limitations and will generally work until the ratio of noise to signal becomes excessive or until the pulses are too smoothed to be detected by the threshold limit in the system.

This exact conditions under which the automatic threshold will give acceptable results needs to be assessed for each specific installation and its particulars. vibro-meter field service personnel are trained to provide such assessments.

In practice then, speed signals can support longer field wiring lengths than their vibration or combustion signal counterparts. This allows to locate speed controllers, or Overspeed detection system (example of the ODS301) farther in a safe location, if the long distance transmission is ok. The same benefit brought by the current modulation, and further by the current+GSI technique apply, as well as in §3.

[12] Abrupt changes in waveforms (such as the edges of rectangular pulses) require very high frequency content to faithfully reproduce because they essentially represent an instantaneous change in amplitude. Theoretically, this requires an infinite frequency response. Practically, a passband that is 10-20X the fundamental frequency is enough to permit a right reconstitution of pulses by an electronic circuitry.
[13] Here we have assumed that the probe is observing a keyway or hole that recesses into the shaft as opposed to a key or protrusion from the shaft, but the principles are the same.

In practice in the field, the issues in scenarios where speed and vibration signals travel long distances together are often: the cross-talk. This phenomenon is encountered when one signal bleeds into another. This can occur when the cables are too close to each other, but can also be caused by other factors.

The transmission from channel to channel occurs due to electromagnetic effects: capacitive coupling and magnetic interferences. This can be particularly problematic when the speed signal has large amplitudes (such as from a proximity probe or even from a magnetic pickup which has an amplitude is proportional to speed) and is transmitted together in the same multi-core cable with small-amplitude signals (such as from an accelerometer measuring in low frequencies). The effect is particularly important when the slope (the "rate-of-change") in the speed pulses are steep. The steepness corresponds to $\frac{dV}{dt}$ (derivative of voltage over time) and is the key parameter controlling the cross-talk. Abrupt shaft discontinuities such as keys, keyways, and holes all result in steep slopes. This cross-talk creates particular difficulties in machinery diagnostics because the pulse frequency coincides with (or -is a multiple of-) shaft rotational speed and will thus give synchronous spikes, and therefore erroneous 1X, as well as harmonics nX vibration readings.

One of the best ways to reduce this – after checking grounds and individual signal shields – is to segregate the speed signals from the vibration signals such that they run in separate cables rather than sharing the same multi-core cable. Choosing also other cable trays, or other terminal blocks in a cabinet, might help reaching the objective. Another solution is to provide low-pass filtering on the speed signal at the source. This can be done by placing a small capacitor in parallel with the device output and tuning it to adjust the cut-off frequency for acceptable results in all cases.

4.2 Frequency modulation with a modulator (quasi-static signals)

When the methods already discussed do not deliver satisfactory results, a frequency modulation (FM) scheme can represent a highly effective way to transmit signals over very long distances. This is because even if the signal amplitude is altered, only the carrier frequency must be preserved in an FM scheme to accurately provide the value. Thus, an FM scheme is far more tolerant to amplitude degradation and distortion than a non-FM scheme.

As seen with the example of speed values (refer to §4.1), this technique can also be effective for long-distance transmission of quasi-static signals such as load, pressure, temperature. While the proportional 4-20mA method is typically used for transmission of such signals over large distances, even this method sometimes reaches its limit when noise begins to overwhelm the signal, whether from electromagnetic interference or from improper shielding/grounding.

The quality of the result depends essentially on the electronic circuitry or the algorithm that is used at the processing end to detect the pulses. Older devices used fixed-level threshold on the signal via a classical "Schmidt trigger" method. Hysteresis is required in this method to avoid double (or multiple) counting of pulses. However, this lacks the flexibility necessary to accommodate the signal attenuation that inevitably occurs with long cables. Acceptable results instead require improved algorithms or devices with both hysteresis and auto-ranging, such as in the VM600-MPC4. For example, this may be the best technique with submerged cable (refer to §1.2). Ultimately, the accuracy and response time can be better than with purely analogue transmission under the same conditions. However note that the frequency limit is, in all cases, much lower than the lowest limit of the carrier frequency used for the modulation and transmission.

The resulting signal can go directly to a frequency channel (*i.e.*, a speed input channel) and the result expressed in the appropriate engineering unit, thanks to the flexibility of the software. The VM600 systems have often spare speed channels that are used to this aim.

An example of an FM transmission scheme is provided in Reference [2] where the 4-20 mA signal is modulated using a [200 – 1000 Hz] frequency span (4mA = 200 Hz, 20 mA = 1000 Hz).

In this example, accuracy, signal-to noise ratio, and response time of the measurement is superior when using an FM transmission scheme into a tachometer channel than when using the raw analogue 4-20 mA signal into a quasi-static channel.

4.3 Transmission with modulator and demodulator

This same frequency modulated signal can also be demodulated by a mirroring device at the signal's destination, transforming the frequency signal back into a conventional analogue signal which can then be input to a conventional dynamic or quasi-static monitoring channel. While the accuracy may not be as good, it will often be acceptable – and may indeed be a viable option for reducing the noise on a long distance transmission.

Of course, in the modulating device, the frequency span must be adapted to the frequency band one wants to analyse. The carrier frequency must always be at least more than twice that of the highest frequency one wishes to analyze. If this is not observed, aliasing will occur and this may lead to erroneous machinery diagnostic conclusions.

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CONCLUSION

Transmission over long field wiring distances can be not detrimental to signal quality, as we can design the systems accordingly through judicious selection of components and cables. Calculations guide the designer to select the right technique and the right cable. Of course all choices are a compromise quality/cost.

For a given combination of distance and bandwidth, if classical voltage-mode signal transmission gives unsatisfactory results, then current-mode transmission will improve the performance. The improvement in frequency band and distance depends on the cable effective capacitance, and the system impedances.

However, if simply employing a current-mode transmission scheme is not sufficient, or if there are concerns with the ground difference between the system and the machine, then a vibro-meter GSI device is recommended, as this will bring further improvement as seen in Figure 7.

While speed signals have numerous issues that must be carefully addressed in design of overspeed, phase reference, and speed control systems, they can often be reliably transmitted over longer distances than conventional dynamic signals such as vibration and dynamic pressure provided appropriate precautions are taken.

As was also shown, even quasi-static signals such as process variables that would normally use conventional 4-20mA transmission can benefit from an FM scheme when very long distances are involved. Such FM signals can be brought directly into spare tachometer channels such as those in our VM600 MPC4^{Mk2} cards.

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Meggitt pioneered high performance sensing and condition monitoring solutions for extreme environments. After working with the world's turbine manufacturers for more than 60 years, Meggitt through vibro-meter portfolio remains master of all aspects of the condition monitoring and machinery protection disciplines.
From high performance sensing, data acquisition and management to the high speed digital networking and the signal processing algorithms that can deliver diagnostics for prescriptive maintenance solutions.

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