AND8466/D

NCS5650 PLC Filter Design

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APPLICATION NOTE

Introduction

Power line communications (PLC) has existed for some time since its introduction in automatic meter reading (AMR) as one of PLC's first applications. Since electrical outlets are ubiquitous throughout the home and office, power line communication is an optimal solution to provide communications between a residential or industrial client and a power distributor. Load control, lighting control, and smart homes are a few additional types of applications. However, providing data communication over the power lines may be difficult due to the power line network being an extremely noisy environment.

The electrical grid of a home or office presents several challenges to the system designer for several reasons. The frequency response of the electrical grid from home to home will vary greatly due to the various of stubs and terminating impedances. This changing impedance is compounded even further as it will also vary in time with the addition of these devices plugged into the electrical grid when turned off and on. Noise sources must also be considered; typical noise sources include brush motors, halogen lamps, and switching power supplies inject noise into the power line. All three hazards provide a difficult data transmission medium to provide reliable data.

ON Semiconductor provides a system level solution to help overcome these issues in PLC applications. The AMIS−49587 PCL carrier modem coupled with the NCS5650 high voltage, high current amplifier are specifically designed for AMR and other PLC oriented applications. This application report will review the CENELEC transmission and disturbance requirements for PLC and how to design ON Semiconductor's NCS5650 PCL line driver to interface into the electrical mains to ensure proper data transmission.

CENELEC Requirements for Power Line Communication

The European regulatory committee responsible for allocating the communication requirements is the Comité Européen de Normalisation Électrotechnique[1] or CENELEC. CENELEC provides five different frequency bands, and maximum transmission and disturbance levels when transmitting data over power lines. Table 1 lists the frequency bands regulated by CENELEC and Table 2 lists the maximum transmission and disturbance amplitudes for a specific frequency band. Figure 1 is the CENELEC transmission and disturbance mask which graphically illustrates the maximum amplitudes for transmitted signals and disturbance signals; e.g., 2nd and 3rd harmonic content, in the CENELEC A−band.

Table 1. (source CENELEC EN 50065−1)

Band	Frequency range	Purpose
	3 kHz − 9 kHz	Electric distribution company use
A	9 kHz − 95 kHz	Electric distribution company use and their licenses
B	95 kHz − 125 kHz	Consumer use with no restrictions
C	125 kHz − 140 kHz	Consumer use only with media access protocol
D	140 kHz − 148.5 kHz	Consumer use with no restrictions

Table 2. (source CENELEC EN 50065−1)

Frequency Range	Maximum Transmission Level	Maximum Disturbance Level
3 kHz − 95 kHz	134 dBμV −120 dBμV	89 dBμV − 75.5 dBμV
95 kHz − 148.5 kHz	116 dBμV	75.5 dBμV − 65.97 dBμV
95 kHz − 148.5 kHz	116 dBμV	75.5 dBμV − 65.97 dBμV

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July, 2010 − Rev. 0

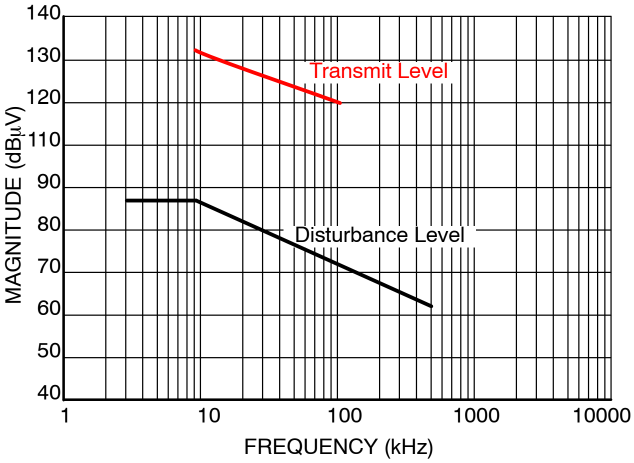

Figure 1. CENELEC A−Band Transmission and Disturbance Mask

There are five different sub−bands in the frequency range allocated by CENELEC. The first two sub−bands according to Table 1 are limited to utility providers and the remaining three are reserved for the customers of the same utility providers.

We will continue with a review of analog filters before delving into the design of the NCS5650 filter topology so the reader may be familiar with the critical filter parameters for a good design.

Review of Analog Filters

Low Pass Filter

Low pass filters, or any filter for that matter, can be solely constructed with passive components or in conjunction with active devices such as operational amplifiers. The text book low pass filter, also known as an integrator, is illustrated in Figure 2, and Figure 3 is a practical active low pass filter.

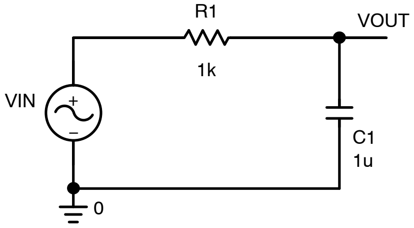

Figure 2.

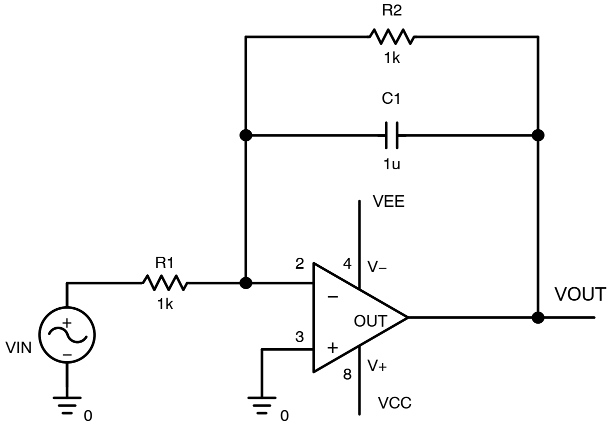

Figure 3.

The ideal and practical responses for the magnitude and phase of a low pass filter are shown in Figures 4 and 5. The magnitude response is often plotted against radians per second which is then normalized to Hertz for our benefit. The notable regions that are highlighted in Figure 4 are the pass band, stopband, and ripple. The passband is the region of the filter where all frequency content is passed unperturbed. The stopband is the region of the filter where all frequency content is considered to be fully restricted.

The ideal magnitude response of a single pole filter, Figure 4, illustrates a −20 dB/dec role off. The ideal phase response, Figure 5, begins to decrease one decade below f_C reaching 45° at the cut−off frequency and will continue to decrease one decade above f_c ending at 90° of phase shift. Put simply, the magnitude will roll off 20 db/dec and the phase will shift 90° respectively per pole.

The idealized bode response of the filter is drawn using straight line segments for approximation and is very close to the practical response. To illustrate the difference between each response, Equation 1, will be used to calculate the actual response at a given frequency. Using Figure 1 as an example, the gain and phase of the output is determined by Equation 1:

$$A_V = \frac{1}{\sqrt{1 + \left(\frac{f}{f_C}\right)^2}} \angle -\arctan\left(\frac{f}{f_C}\right) \qquad \text{(eq. 1)}$$

It is clear that there are small errors from the idealized bode plots, but for first order analysis they remain very useful to quickly plot the filter’s response.

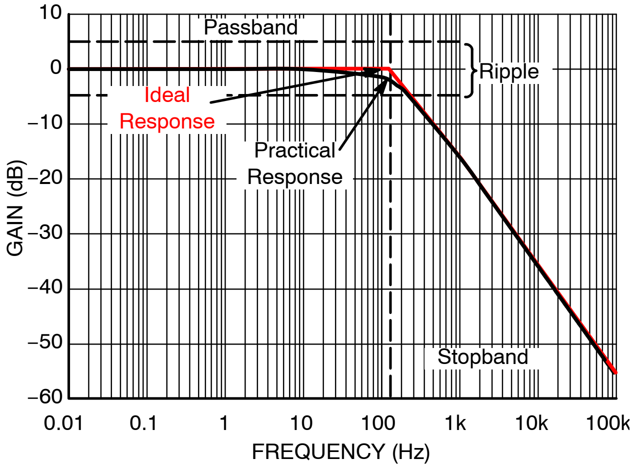

Figure 4. Ideal and Practical Magnitude Response

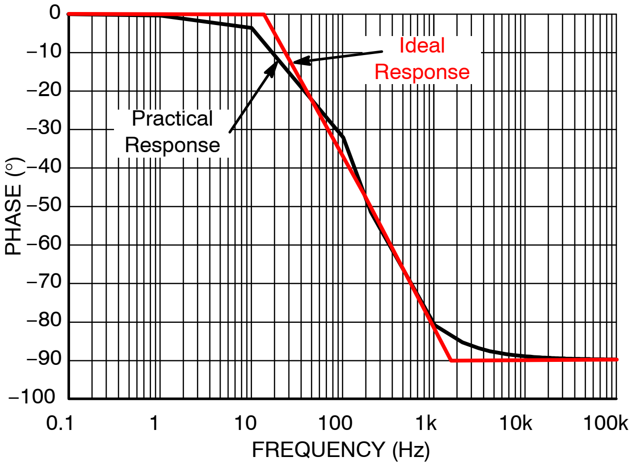

Figure 5. Ideal and Practical Phase Response

The transfer functions of both circuits are different; however, their similar frequency response will be examined. The transfer functions will be manipulated in the frequency domain using the LaPlace transforms for ease of calculation rather than cumbersome differential equations representative of the time domain. For Figure 1 the output transfer function defined by:

$$\left(\frac{V_{OUT}}{V_{IN}}\right) = \left(\frac{1}{(R_1C_1)s + 1}\right)$$

For Figure 2 the transfer function is

$$\left(\frac{V_{OUT}}{V_{IN}}\right) = -\frac{R_2}{R_1}\left(\frac{1}{(R_2C_1)s + 1}\right)$$

For the passive low pass filter, normalizing the transfer function equates the angular frequency[3], ω_n to (1/RC) and this is shown in Equation 2:

$$\omega_n = \frac{1}{RC} \qquad \text{(eq. 2)}$$

Solving for f_c, the cut−off frequency for the passive low pass filter is easily derived to be $f_c = 1/(2*\pi*R_1C_1)$. The cut−off frequency of the active filter will normalize in a similar fashion: $f_c = 1/(2*\pi*R_2*C_1)$. The remaining term is the inverting amplifier gain, $-(R_2/R_1)$.

We will focus on the CENELEC A frequency band as an example; however, the same principles will hold for any of the frequency band ranges listed in Table 1. The frequency range for the CENELEC A band is 9 kHz to 95 kHz. The NCS5650 evaluation board was designed with a 95 kHz cut−off frequency using a multiple feedback topology. In order to meet this design specification, we will review the construction of a 4th order filter from basic filter building blocks, the Butterworth response, and the MFB topology.

Butterworth Frequency Response

The Butterworth filter is one of several types of filter responses available for design; other popular filters include Chebyshev, Elliptic, and Bessel.

The Butterworth filter response is often desired when passband gain is required to be maximally flat or no passband ripple. One of the trade−offs for this extremely flat response is the Butterworth response does not have as sharp a roll off as other filters of the same order. The flat magnitude response of the Butterworth filter is shown in Figure 6. For the NCS5650 demoboard, the Butterworth frequency response was implemented to ensure a flat response; however, a 4th order Butterworth filter is required due to its mild roll off.

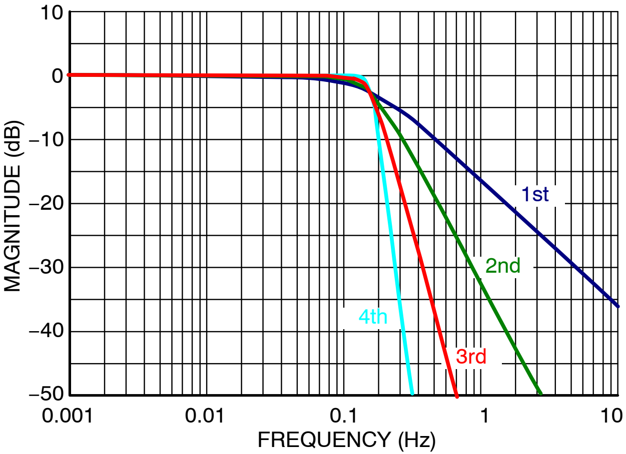

Figure 6. Magnitude responses for 2nd, 3rd, and 4th order filters

Multiple−Feedback Filter Topology

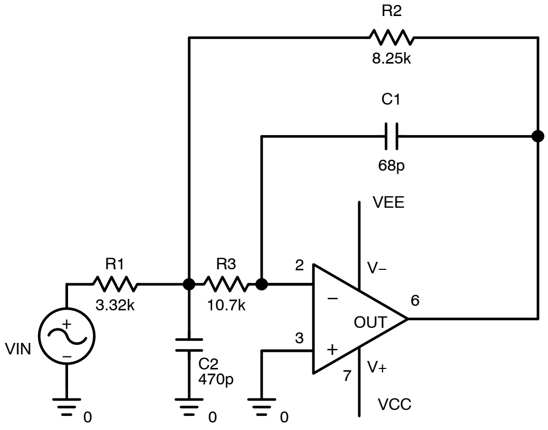

Figure 7. MFB Filter Topology

Multiple feedback (MFB) filters can only be realized with the use of active elements like operational amplifiers. Figure 7 is a typical 2nd order MFB lowpass filter. MFB filters build upon the inverting amplifier configuration and embed and integrator, R_3 and C_1 in this case, within a feedback loop created by R_1, R_2, and C_2. The MFB topology is less prone to errors due to component variations and has robust high frequency response when compared to the Sallen−Key topology[4].

Standard form of the second order equation

The standard form used for transfer functions of second order systems is shown in Equation 3:

$$K(s) = \frac{k\omega_n{}^2}{s^2 + (2\varsigma\omega_n)s + \omega_n{}^2} \qquad \text{(eq. 3)}$$

The coefficient k is the DC gain of the system, ω_n is the undamped natural frequency and the coefficient ζ is the dampening ratio. ζ is the term used often in control theory while Q, or quality factor, is typically used when discussing filters. The relation of Q to ζ is expressed in the following Equation 4:

$$2\varsigma = \frac{1}{Q} \qquad \text{(eq. 4)}$$

So Equation 4 can be re−written as Equation 5 which is often more familiar when designing filters:

$$K(s) = \frac{k\omega_n{}^2}{s^2 + \left(\frac{\omega_n{}^2}{Q}\right)s + \omega_n{}^2} \qquad \text{(eq. 5)}$$

Transfer Function of a 2nd Order MFB Filter

Referring to Figure 7, the transfer function is shown in Equation 5 and is in the standard second order form. Equation 6 can be used directly to calculate the signal amplification of the circuit for a given input signal.

$$K(s) = \frac{\left(-\frac{R_2}{R_1}\right)\left(\frac{1}{C_1C_2R_2R_3}\right)}{s^2 + s\left(\frac{1}{C_2R_1} + \frac{1}{C_2R_2} + \frac{1}{C_2R_3}\right) + \left(\frac{1}{C_1C_2R_2R_3}\right)} \qquad \text{(eq. 6)}$$

Comparing Equation 6 to Equation 3 allows the reader to quickly identify the proper coefficients when designing active filters utilizing the multiple feedback architecture:

$$k = \left(-\frac{R_2}{R_1}\right) \text{ is the DC gain.} \qquad \text{(eq. 7)}$$

$$Q = \left(\frac{R_1\sqrt{R_2R_3C_1C_2}}{C_1(R_1R_2 + R_1R_3 + R_2R_3)}\right) \text{ is the quality factor.} \qquad \text{(eq. 8)}$$

$$\omega_n = \left(\frac{1}{\sqrt{C_1C_2R_2R_3}}\right) \text{ is the undamped natural frequency.} \qquad \text{(eq. 9)}$$

Filter Building Blocks

The approach with active filter design is to use basic filter building blocks. Each section will be a 1st or 2nd order filter block, and to achieve higher order filters 1st and 2nd order filter stages are cascaded as illustrated in Figure 8. When cascading filter blocks each stage requires a frequency scaling factor, FSF, and subsequent Q in order to preserve the overall filter response, and these responses are derived from cumbersome polynomial equations.

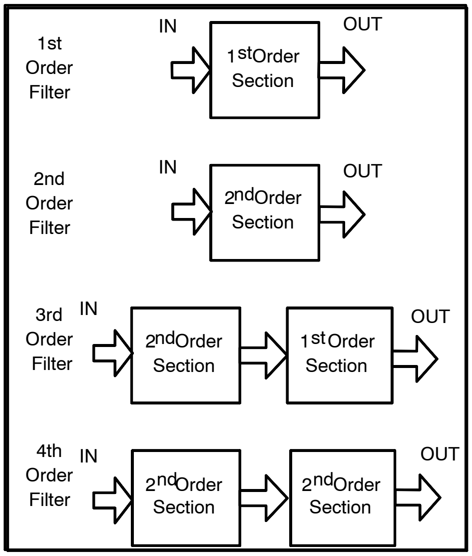

Figure 8. Realization of Higher Order Filters by Cascading Filter Stages

Thankfully there are resources that present look up tables when designing filter circuits rather than dealing with cumbersome polynomial expression. This design note will take the approach of using these classic filter tables that are ubiquitous in analog filter reference design books. The filter tables are often used to reduce the heavy mathematical calculations used to determine the necessary R and C component values for the filter circuit. They serve as a quick design reference ratio once several parameters are chosen beforehand.

Each filter type has its own coefficient table; i.e., Butterworth, Bessel, Chebychev, based on the desired filter order. Table 3 lists the frequency scaling factors and circuit Q necessary for a Butterworth filter to ensure a flat response in the passband up to a 10th order circuit. For those interested in other filter tables additional references are available[6].

Table 3. BUTTERWORTH FILTER COEFFICIENTS

Filter Order	Stage 1		Stage 2		Stage 3		Stage 4		Stage 5	
	FSF	Q	FSF	Q	FSF	Q	FSF	Q	FSF	Q
2	1.0000	0.7071								
3	1.0000	1.0000	1.0000							
4	1.0000	0.5412	1.0000	1.3065						
5	1.0000	0.6180	1.0000	1.6181	1.0000					
6	1.0000	0.5177	1.0000	0.7071	1.0000	1.9320				
7	1.0000	0.5549	1.0000	0.8019	1.0000	2.2472	1.0000			
8	1.0000	0.5098	1.0000	0.6013	1.0000	0.8999	1.0000	2.5628		
9	1.0000	0.5321	1.0000	0.6527	1.0000	1.0000	1.0000	2.8802	1.0000	
10	1.0000	0.5062	1.0000	0.5612	1.0000	0.7071	1.0000	1.1013	1.0000	3.1969

Design Example

The NCS5650 evaluation module is designed to meet the CENELEC A frequency band. In order to achieve the necessary attenuation for the transmit and disturbance mask in the A band, a 4th order filter, multiple feedback, low pass filter is used. Reduced filter orders and other topologies are possible and their design implementation is left as an exercise to the designer.

Figure 9 illustrates the 4th order MFB filter architecture for the NCS5650. Values must be calculated for each stage of the filter which can be a difficult process even with the simplifications previously given.

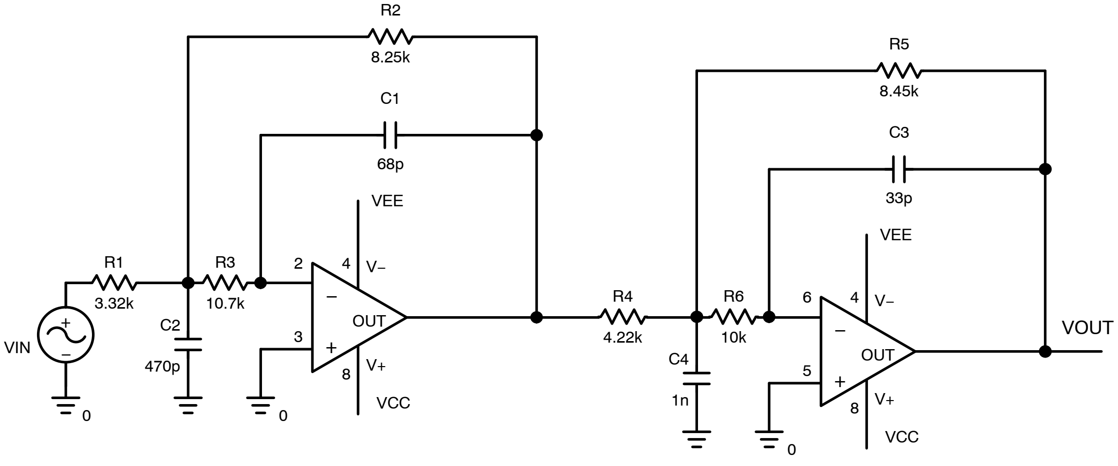

Figure 9. NCS5650 4th Order Filter Implementation

As previously mentioned the use of filter tables will be used in conjunction with several known variables and derived equations beforehand to help the ease of calculation. A desired circuit gain and cutoff frequency should be chosen before design begins and the use of Equations 10 and 11 will facilitate the derivation of component scaling coefficients m and n.

The variables m and n represent resistor and capacitor scaling factors respectively. Assuming that $R_2 = R$, $R_3 = mR$,

$C_1 = C$ and $C_3 = nC$ upon inspection of Equations 8 and 9 we can arrive at Equations 10 and 11 as illustrated below:

$$f_C = \frac{1}{2\pi RC\sqrt{mn}} \qquad \text{(eq. 10)}$$

$$Q = \frac{\sqrt{mn}}{1 + m(1 - K)} \qquad \text{(eq. 11)}$$

The target gain, cut−off frequency, and seed value for R_2 are:

$$A_{VDC} = -2.5$$

$$f_C = 95\ \text{kHz}$$

$$R_2 = 8.25\ \text{k}\Omega$$

With the above equations and known resistance, the component scaling coefficients m and n can be calculated in order to determine R_1, R_3, and C_2.

Solving for m and n will require Equations 10 and 11, Q from the filter table for the given filter order and stage, target gain, cut−off frequency, and the use of simultaneous equations.

Using Equation 10, the variable $\sqrt{mn}$ is isolated and substituted into Equation 11 to and determine the resistor scaling coefficient, m. After m is determined, its value can be substituted back into Equation 10 to solve for the capacitor scaling coefficient, n. These steps are briefly shown below:

$$Q = \frac{\left(\frac{1}{2\pi f_C RC}\right)}{1 + m(1 - k)}$$

$$0.5412 = \frac{\left(\frac{1}{2\pi \times 95\ \text{kHz} \times 8.25\ \text{k}\Omega \times 68\ \text{pF}}\right)}{1 + m(1 - (-2.5))}$$

m = 1.2908

Substituting m back into Equation 7.1 will yield a capacitor scaling coefficient of: n = 6.9058

Recalling $R_2 = R$, $R_3 = mR$, $C_1 = C$ and $C_3 = nC$ it is now determined that $R_3 = 10.7\ \text{k}\Omega$ and $C_2 = 470\ \text{pF}$.

R_1 is calculated from Equation 6.1 and is determined to be 3.32 kΩ.

Repeating the same calculations for the second stage remembering to use the appropriate Q for the second stage in a 4th order filter, 1.3065, will determine the necessary component values. For brevity these are already provided below:

$R_4 = 4.22\text{k}$, $R_5 = 8.45\text{k}$, $R_6 = 10\text{k}$, $C_3 = 33\text{p}$, $C_4 = 1000\text{p}$.

Summary Design Steps:

The steps required to begin filter design are summarized below.

- Circuit gain, cut−off frequency, and component values R_2 and C_1 are chosen before design.
- C_1 is restricted with the rule of thumb $10\ \text{pF} < C_1 < 100\ \text{pF}$.
- Use Equation 10 and solve for the variable $\sqrt{mn}$ which will be substituted into Equation 11.
- Solve for the resistor scaling coefficient, m and substitute its value into Equation 10.
- Solve for the capacitor scaling coefficient, n.
- Solve for R1 using the known circuit gain and R_2.
- Given $R_2 = R$, $R_3 = mR$, $C_1 = C$, $C_2 = nC$ solve for R_3 and C_2.
- Repeat the process for the succeeding filter stages using the appropriate FSF and Q from Table 3.

Following the above process will ease the difficulty of calculating component values for the filter stages.

Spice Simulation:

Substituting the component values that were previously calculated back into Figure 9, the simulation results Figure 10 show excellent passband flatness and a cut−off frequency of 95 kHz:

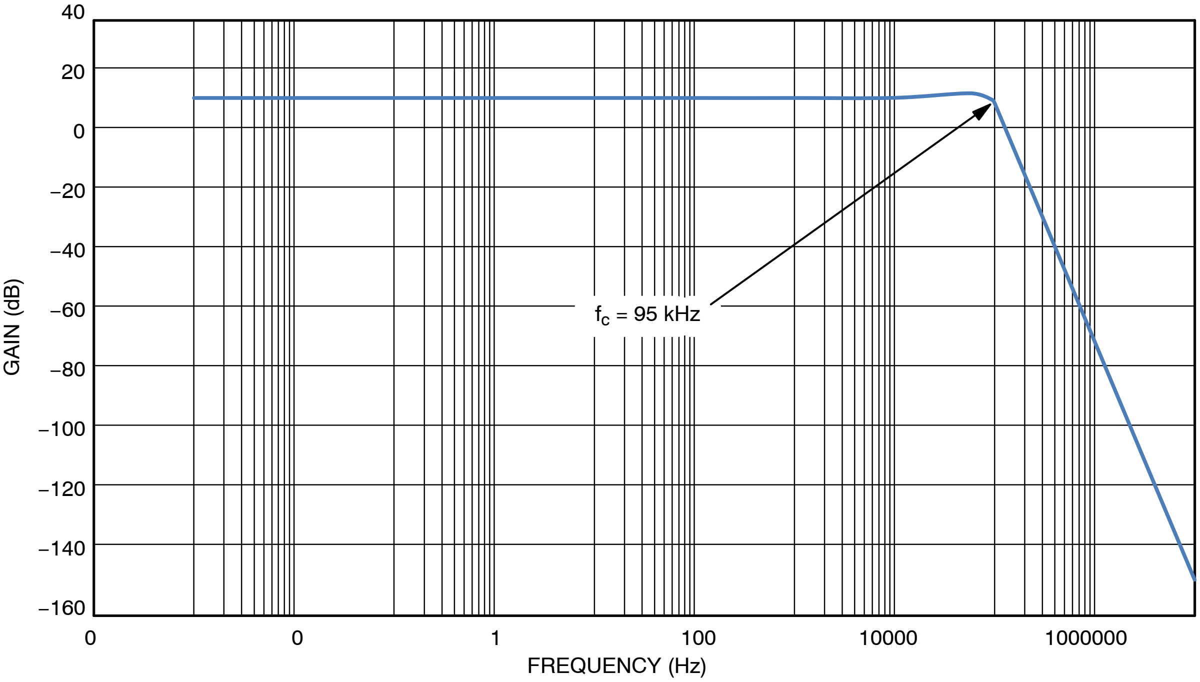

Figure 10. Gain and Phase of a 4^{th} Order MFB Filter

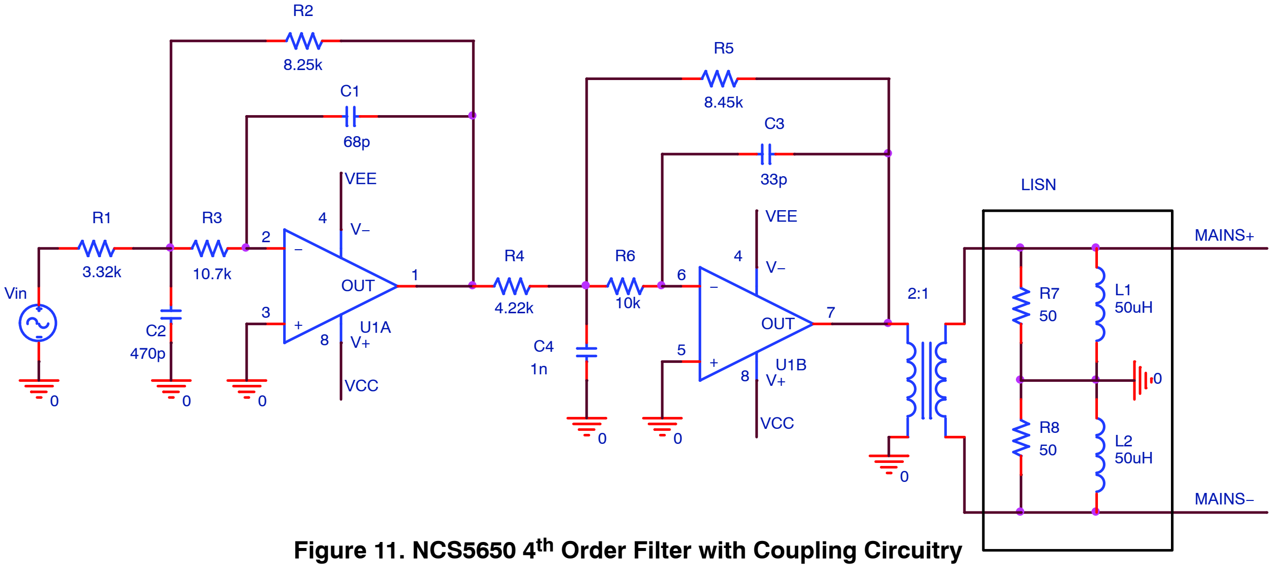

Figure 11. NCS5650 4^{th} Order Filter with Coupling Circuitry

The AMIS−49587 modem uses 63.3 kHz and 74 kHz for mark and space frequencies when utilizing frequency shift keying (FSK). Referencing back to Tables 1 and 2, the maximum signal level for the A band is 134 dBμV at 9 kHz and 120 dBμV at 95 kHz. The maximum disturbance levels for the 2^{nd} harmonics are 67.3 dBμV for 63.3 kHz and 66 dBμV for 74 kHz.

CENELEC also calls for a line impedance stabilization network (LISN)7 when devices are coupled to the electrical mains. The purpose of the LISN is to provide a defined impedance across the electrical mains. The LISN together with the AMIS−49587 and NCS5650 filter circuit provide the essential circuit to meet the requirements called for by CENELEC.

The addition of a 2:1 isolation transformer and the LISN, Figure 11, further reduces the measured transmission and disturbance levels as shown in Figure 12. Since the measurement is made after the LISN circuitry the fundamental will be 122 dBμV. The 10 dB attenuation from the 4^{th} order filter and additional 60 dB of attenuation from the AMIS−49587 modem at the second harmonic ensures the disturbance level is 52 dBμV at 126.6 kHz. This is sufficient to meet the CENELEC specification.

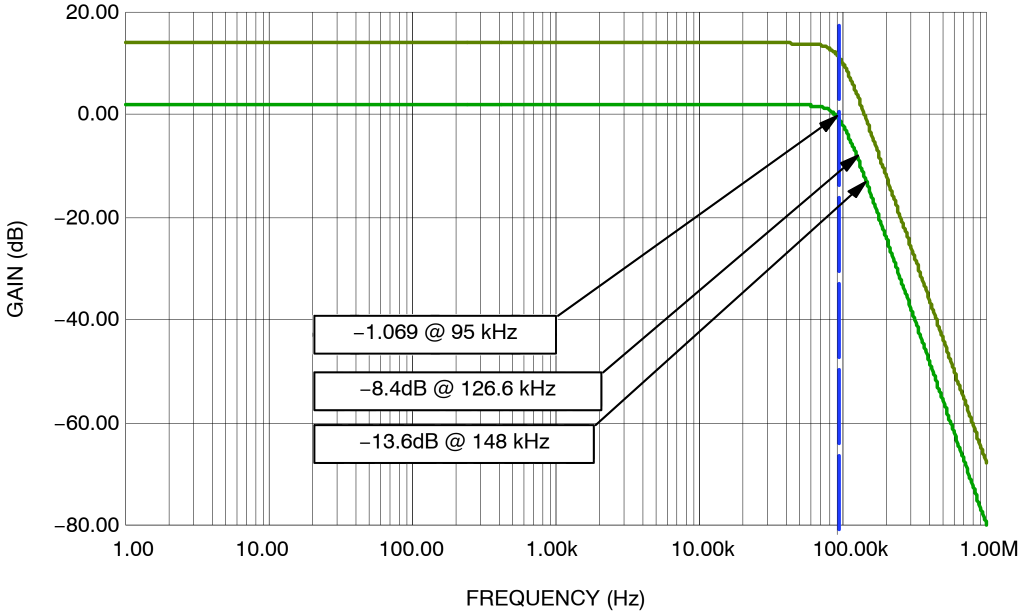

Figure 12. Frequency Response of NCS56504th Order Filter, Coupling, and LISN

Limitations:

The obvious limitation to the component values is finding a standard value for the resistors and capacitors as close to the calculation as possible. Using 1% tolerances from the E96 family will help provide more options to the available practical values.

Filter sensitivity to component values is another concern; although, the multiple feedback architecture is less sensitive than its Sallen−Key counterpart. The mathematics behind component sensitivity becomes very cumbersome and is beyond the scope of this application note, but the engineer should be made aware of its effects.

The component scaling values m and n can theoretically be of any value and at least one may be the same value for each 2nd order stage; however, this can lead to additional peaking when combined with amplifier gain. It is advised that the component scaling values are different for each stage.

Finally, the measurements above used an ideal transformer so there is no consideration to the frequency response of the transformer due to saturation, leakage inductance, or capacitance winding.

Summary:

This design note reviewed the CELENEC requirements for transmission and disturbance levels onto the electrical mains by analyzing the necessary filter design requirements for the NCS5650 to work in conjunction with the AMIS49587 and LISN. Other filter topologies, stage orders, and coupling networks – direct or transformer coupled are possible, but are left for an exercise for the design engineer.

References:

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2. Introductory Circuit Analysis – Boylestad, Prentice Hall; 9 edition (August 16, 1999) pg. 935
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