FREQUENCY RESPONSE OF DIGITAL LOCK-IN TECHNIQUE FOR POWER-LINE INTERFERENCE EXTRACTION

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Power-line interference is a common problem in almost all biosignal acquisition applications. Recently a smart approach for PL suppression, called lock-in technique, was developed. This paper discusses the behavior of open-loop lock-in technique in frequency domain. It shows that the low-pass transfer function of the used filter is converted to high-pass function by a simple subtraction from unity and then it is transposed in two sidebands around PL frequency. Thus, the flatness roll-off characteristic of the used low-pass filter is very important for achievement of final rippleless frequency response of the lock-in filtering approach. A simple digital low-pass filter is proposed to be used in cases when a maximally-flat frequency response is needed.

Keywords: power line interference, mixer, synchronous demodulator, phase-sensitive demodulator, lock-in amplifier

1. INTRODUCTION

Power-line (PL) interference (hum) is a common problem in almost all biosignal acquisition applications. Because the body serves as a capacitively coupled antenna, a part of the picked up PL interference currents traverses the electrodes and produces a common mode voltage over an amplifier common mode input impedance. At the amplifier output some AC noise remains as a consequence of electrode impedance imbalance and/or due to the finite value of the amplifier CMRR [1], even when special signal recording techniques are applied (shielding, driven right leg, body potential driving, etc.). A further reduction of the interference should be implemented by either post-digital or post-analog filters.

Recently a smart approach for PL suppression, called lock-in technique, was developed [2, 3].

This paper discusses the behavior of the open-loop lock-in technique in frequency domain. It shows that the low-pass transfer function of the used filter is converted to high-pass function by a simple subtraction from unity and then it is transposed in two mirrored sidebands around PL frequency. Thus, the flatness roll-off of the used low-pass filter is very important for achievement of final rippleless frequency response of the whole lock-in filtering approach.

2. OPEN-LOOP LOCK-IN CONCEPT

The open-loop lock-in concept for PL interference extraction [2, 3] is redrawn in Fig. 1. It consists of two square wave mixers: in-phase mixer M_2 and quadraturephase mixer M_1. After low-pass filtering, the average values at M_2 and M_1 outputs are:

$$V_{Re} = V_m \frac{2}{\pi}\cos\theta, \qquad V_{Im} = -V_m \frac{2}{\pi}\sin\theta \qquad (1)$$

Using V_{Re} and V_{Im} from (1), the PL interference V_I can be synthesized by a sum of sinusoidal and cosinusoidal waveforms according the equation (2) [1, 2]:

$$V_I = V_m\sin(\omega t+\theta) = \frac{\pi}{2}V_{Re}\sin(\omega t) - \frac{\pi}{2}V_{Im}\cos(\omega t) \qquad (2)$$

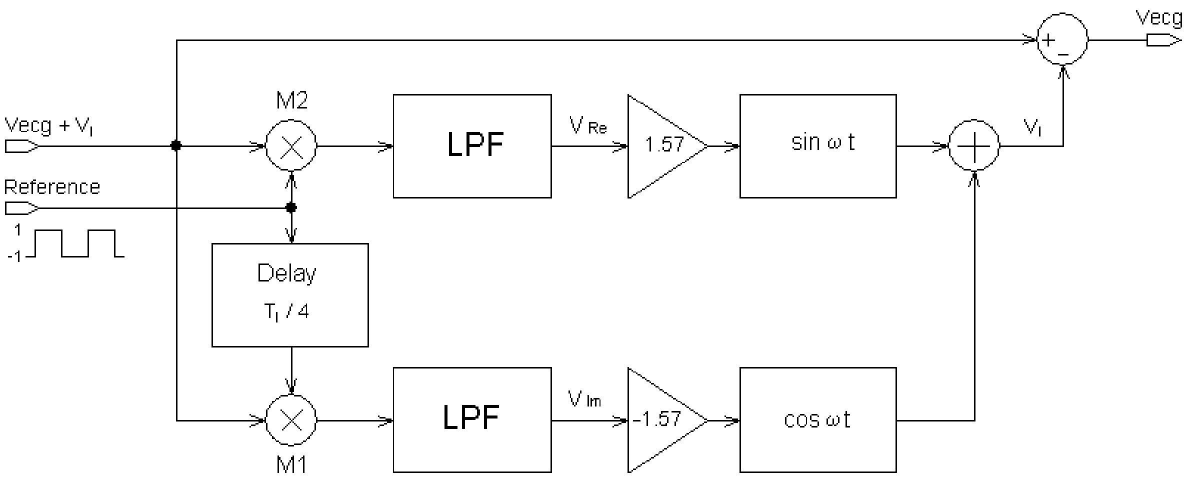

Fig. 1 Open-loop concept for PL interference extraction

In a nutshell, the open-loop lock-in filtering approach (see Fig. 1) first measures the amplitude and phase of PL frequency, then generates a complex sine wave with the same parameters, and finally subtracts the generated sine wave from the incoming data stream.

3. FREQUENCY RESPONSE

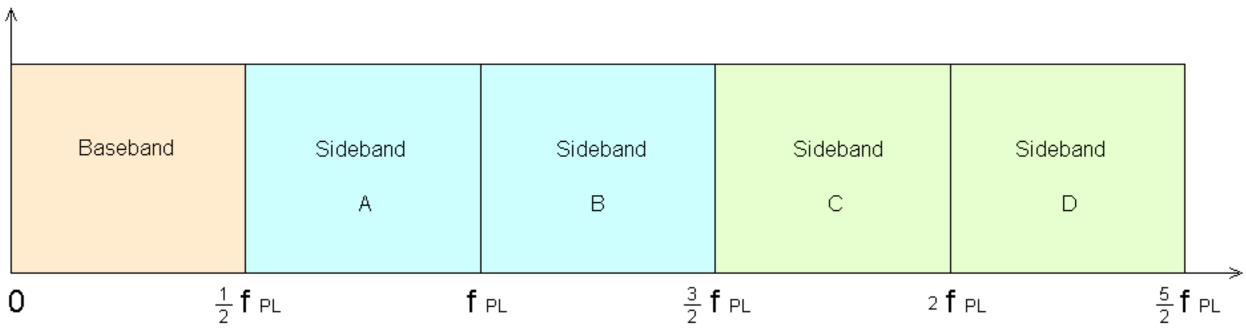

Fig. 2 Baseband and two sidebands around F_{PL} and $2f_{PL}$

Let us consider that the input signal has constant spectral density as shown in Fig. 2, and also that the frequency of internally generated reference f_{REF} is equal to the PL frequency f_{PL}. The spectrum from zero frequency (DC) to $2.5f_{PL}$, can be separated into five parts: baseband from DC to $f_{PL}/2$, two sidebands: A and B, around PL frequency, and two sidebands: C and D, around the second harmonic of PL frequency.

The basic trigonometry says that when two sinusoidal signals are multiplied, their product is represented by the sum of two new sinusoids, one with sum and the other with difference frequencies. The new sinusoids have amplitude equal to the half of the product of amplitudes of their originals.

According to the above statement, for sidebands A and B, mixers M_1 and M_2 (see Fig 1) serve as a demodulator (down-converter) and their spectral components are shifted into baseband due to frequency differencing with f_{REF}. The same stectra A and B, are shifted also into sidebands C and D due to summing with f_{REF}.

At the same time, for frequency components found into baseband, M_1 and M_2 serves as a modulator (up-converter) and the baseband is shifted into the sidebands A and B.

For simplicity, our mixers M_1 and M_2 work with square wave reference, with ±1V swing. This reference can be expressed as an infinite Fourier series, containing only odd harmonics:

$$V_{REF} = \frac{4}{\pi}\sum_{n=1}^{\infty}\frac{\sin((2n-1)\omega_0 t)}{2n-1} = \frac{4}{\pi}\left(\sin\omega_0 t + \frac{1}{3}\sin 3\omega_0 t + \frac{1}{5}\sin 5\omega_0 t + ...\right) \quad (3)$$

It is clear that the same things, as discussed above will be performed for each harmonic of (3) and the sidebands around these harmonics will be transformed into the baseband and will introduce an error and inaccurate f_{PL} synthesis. The odd harmonic response is the one and only one drawback of square wave mixing. For reducing of this error in cases when the lock-in procedure is running repeatedly, the procedure should start from the highest harmonic selected for removal.

Let's come back to the M_1 and M_2 outputs. After low-pass filtering the outputs of M_1 and M_2 contain a slowly varying voltage close to DC, and this voltage serves as an amplitude of artificially generated sine wave with PL frequency. In other words this voltage is appeared multiplied by a sinusoidal signal with amplitude of unity (1V) and frequency f_{PL}. This multiplication performs a new mixing, up-conversion, and shifts the filtered baseband again into sidebands A and B, i.e. the frequency response of the used low-pass filter is shifted in two mirrored sidebands around PL frequency.

Because the generated sine wave is subtracted from the input data, the filter transfer function is converted from low-pass to high-pass by a simple subtraction from unity.

The conclusion is that the lock-in frequency response depends mainly on the frequency response of the used low-pass filter. If we want to see the final lock-in frequency response, the transfer function of the used low-pass filter should be converted to high-pass, by a simple subtraction from unity and then this characteristic should be shifted in two mirrored sidebands around the PL frequency.

The low-pass filter described in [2, 3] consists of two cascaded averagers, the first one with 1 PL period (20ms) averaging time, and the second one with 10 PL periods (200ms) averaging time. The low-pass response and its high-pass derivative response of such cascaded averager, achieved by subtraction of filtered low-pass signal from the input, are shown in Fig. 3.

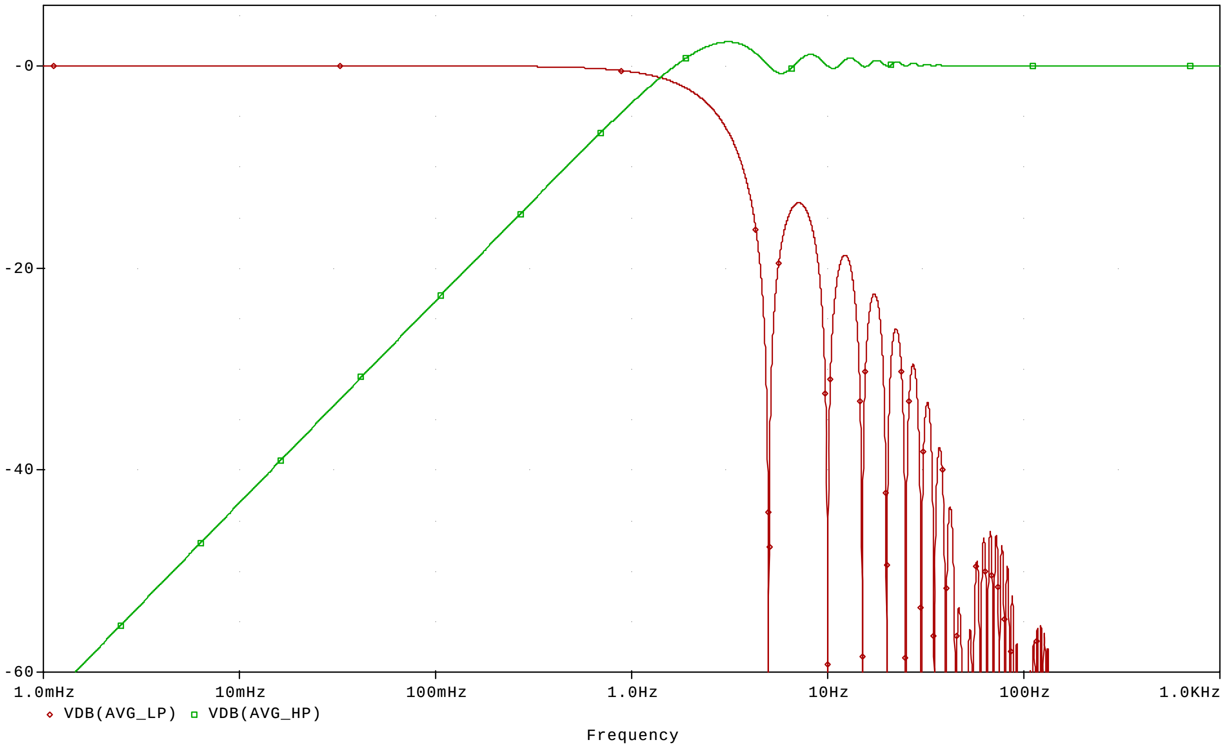

Fig. 3 Frequency response of 1 PL – 10 PL periods moving-average filter together with its high-pass derivative

As was discussed, this high-pass characteristic will be shifted and mirrored in two sidebands around PL frequency. So the final lock-in characteristic will have decaying ripples with first maximum of 2.4dB.

For rippleless lock-in response the used low pass filter should have a flat roll-off, with no lobes or zeros in its stop-band.

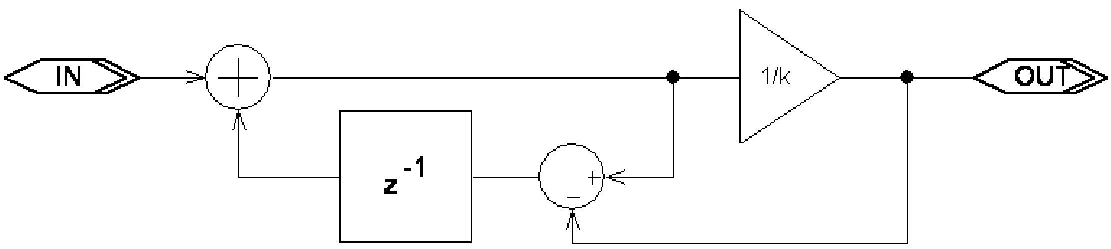

Fig. 4 Simple digital equivalent of a first order analog filter (lossy integrator)

A simple digital equivalent of a first order analog filter (lossy integrator) is proposed to be used in cases when a flat lock-in frequency response is needed. The filter structure is shown in Fig. 4. It has the following transfer function:

$$T(z) = \frac{1}{k-(1-k)z^{-1}} \tag{4}$$

The filter 3dB frequency, as in its analog prototype, is:

$$f_{3dB} = \frac{1}{2\pi\tau} \tag{5}$$

Here $\tau = \frac{k}{f_s}$, and f_s is the sampling frequency.

For simple calculations, the coefficient k should be selected to be a multiple of 2. In our case k = 256.

The quality factor Q of the lock-in approach depends on k and can be approximated as:

$$Q = \frac{f_{PL}}{2f_{3dB}} = \frac{50\pi\kappa}{f_s} \tag{6}$$

If k = 256, and f_s = 2 kHz, then Q = 20.

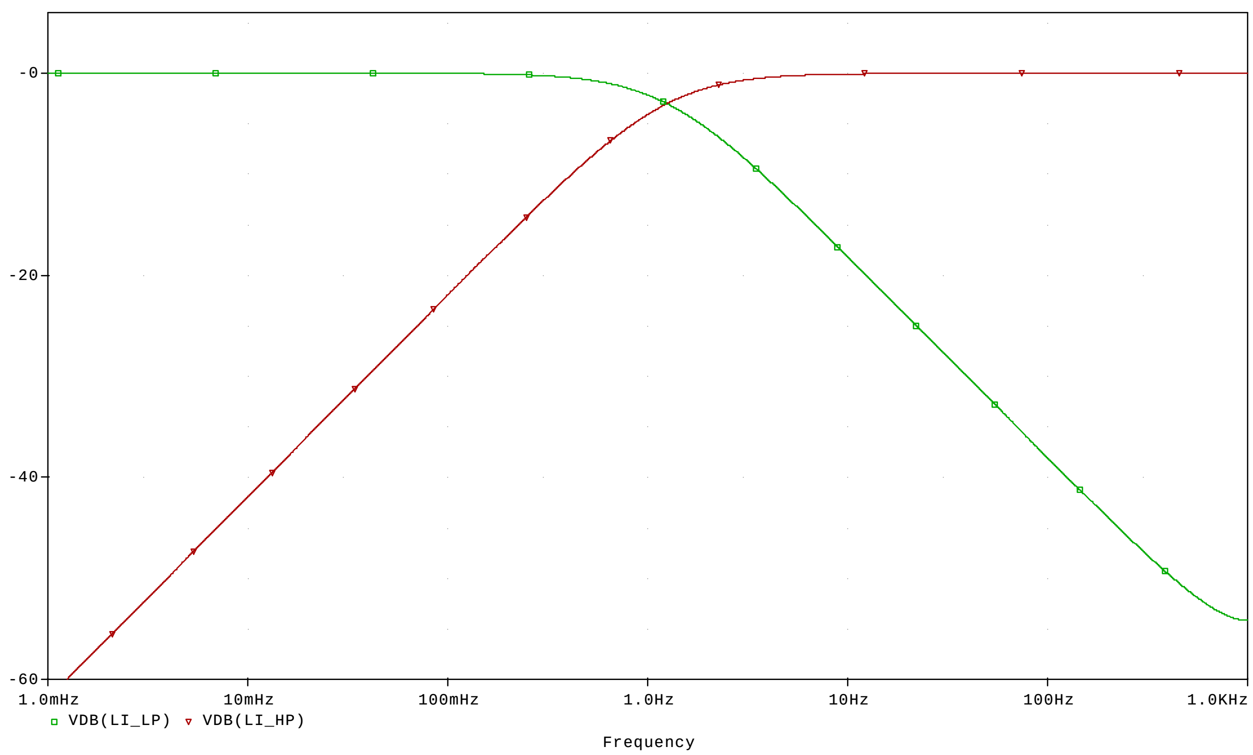

Fig. 5 Frequency response of the low-pass filter from Fig. 4 for k = 256, and its high-pass derivative achieved by subtraction of the low-pass filtered signal from the input

The simulated frequency response of the low-pass filter from Fig. 4 and its high-pass derivative, when k = 256, are shown in Fig 5.

The final lock-in frequency response, for an averager and a lossy integrator cases, will look like this shown in Fig. 6. As discussed above, the achieved high-pass derivative characteristics are shifted and mirrored around PL frequency.

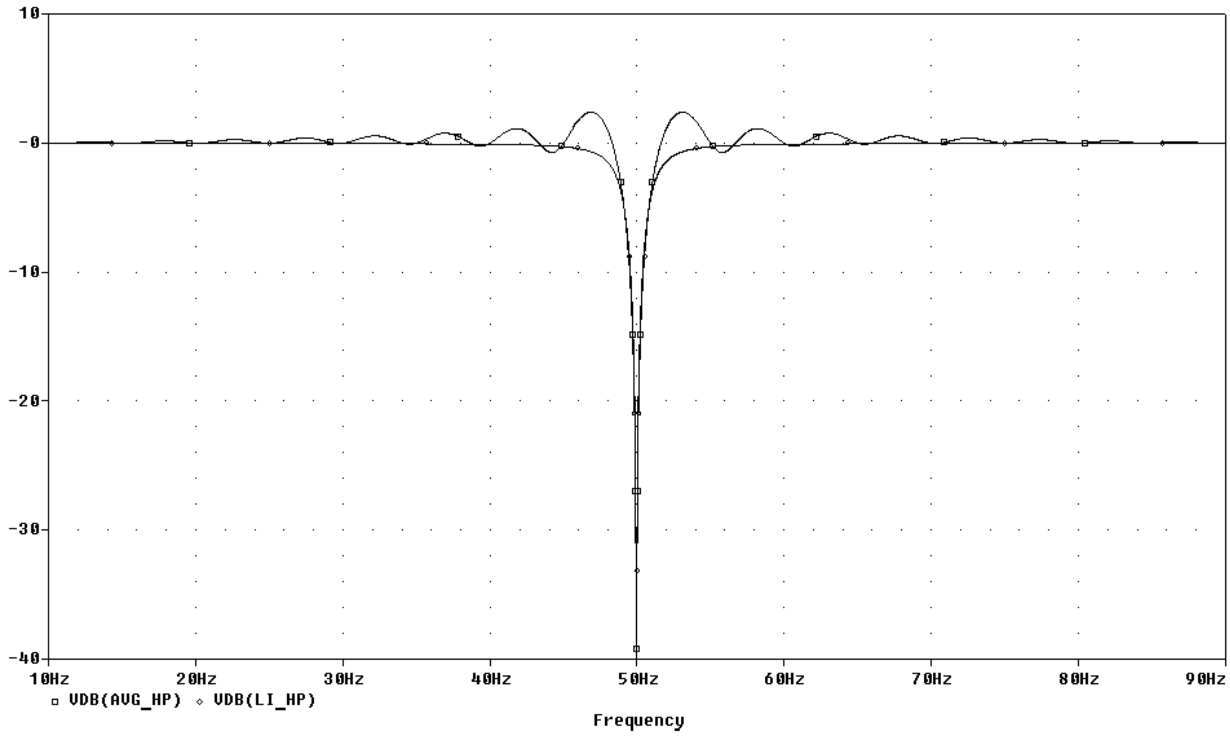

Fig. 6 Lock-in frequency response in cases of an averager and a lossy integrator as low-pass filters

4. CONCLUSION

The lock-in frequency response depends mainly on the frequency response of the used low-pass filter. For simple representation of final lock-in response in frequency domain, the low-pass transfer function of the used filter should be converted to high-pass, by simple subtraction from unity and then this characteristic should be shifted in two mirrored sidebands around the PL frequency.

In cases when rippleless frequency response is needed, the used low pass filter should have a flat roll-off, with no lobes or zeros in its stop-band area.

In such cases the simplest solution is to use a digital equivalent of a first order analog filter (lossy integrator) instead of averager.

5. REFERENCES

[1] Huhta C, Webster J *60-Hz interference in electrocardiography*. IEEE Trans. Biomed. Eng. 20, pp. 91-101, 1973

[2] Dobrev D et al *Digital lock-in techniques for adaptive power-line interference extraction.* Proceedings of the 16th international scientific and applied science conference: Electronics ET2007, Book 2, pp. 9–14, 2007

[3] Dobrev D et al *Digital lock-in techniques for adaptive power-line interference extraction.* Physiol. Meas. 29, pp. 803–816, 2008