# DIGITAL FLICKERMETER REALISATIONS IN THE TIME AND FREQUENCY DOMAINS

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**Abstract**
Flicker monitoring is currently made with a special purpose analogue based instrument specified by the International Electrotechnical Commission. With the current emphasis on digital data acquisition this paper presents time and frequency domain implementations of the flickermeter capable of integration with all-purpose data acquisition systems.

## 1. INTRODUCTION

Limiting flicker annoyance to humans or affecting the operation of sensitive equipment requires objective measurements capable of identifying the flicker sources.

The International Electrotechnical Commission (IEC) has published a standard for flicker measurement [1] which evolved from a proposal for a flickermeter by the Union Internationale d'Electrothermie (UIE) [2]. The standard provides design specifications for an analogue flickermeter but digital realisations are also acceptable. Although digital flickermeter alternatives have already been proposed [3-6], the guidelines for their implementation are vague and compliance with the IEC standard is not given in some cases.

The addition of flicker measurement to CHART, an existing [7] digital data acquisition system, motivated the development of a fully digital design based on the analogue IEC flickermeter specification.

The required analogue-to-digital filter conversions involved are straightforward except for the conversion of the weighting filter, which is given special consideration in this paper.

An alternative flickermeter based on the Discrete Fourier Transform (DFT) is also described which allows for arbitrary shaped weighting filter responses. Multirate processing and specialised partial FFTs are used throughout to ensure maximum computational efficiency.

## 2. DIGITAL, TIME DOMAIN BASED FLICKER-METER

The flickermeter design, except for the statistical evaluation block, is based on the IEC analogue technique described in appendix. However a DSP implementation requires a fully digital realisation.

The s-transfer functions for all filters have been specified by the IEC. The corresponding z-transfer functions can be obtained by standard analogue to digital filter conversion algorithms or designed directly to specification. Figure 1 shows a block diagram of the developed digital flickermeter.

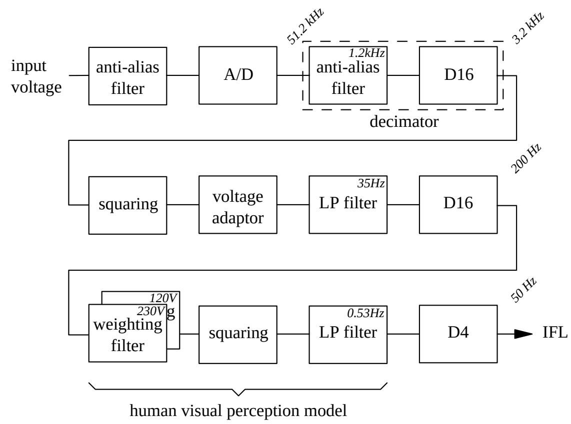


Figure 1: Block diagram of the digital flickermeter (statistical evaluation excluded). Sample rates and filter 3 dB frequencies are indicated in italics.

For the CHART implementation an anti-aliasing filter and a 16-bit analogue-to-digital converter are situated at the front end. From there, fibre optic cables feed the converters at a constant rate of 51.2 kHz digitised samples to the DSPs. For power system monitoring applications this sampling frequency is relatively high, but oversampling has been chosen to simplify the construction of the analogue anti-aliasing filter and also to improve the effective signal-to-noise ratio [8]. The sampling frequency can then be reduced digitally to an appropriate value depending on the application.

The small bandwidth required for flicker measurements allows for an initial sample rate reduction, where 3.2 kHz was chosen giving a Nyquist rate of 1.6 kHz which is more than twice the required 700 Hz [1]. Oversampling slightly allows for a wider transition band in the anti-aliasing filter design which will result in a lower filter order and avoids aliasing effects, due to the squaring operation (directly following the decimator) which essentially doubles the signal bandwidth.

After sample rate decimation the signal is processed through the actual flickermeter. The function blocks shown in Figure 1 correspond to those of the analogue, except that the demodulation high pass filter has been made redundant by fine tuning the weighting filter so that it rejects d.c. completely.

The voltage adaptor originally normalises the signal to the (1 min) average RMS value of the voltage. Since the squaring operation is needed in calculating the RMS voltage, the squaring block was shifted to precede the voltage adaptor which now normalises the signal to the squared average RMS value instead and as a consequence there is no need to compute the square root. For the low pass filter incorporated in the voltage adaptor a $1^{st}$ order Butterworth design was chosen. The corresponding discrete-time filter can be obtained by the bilinear transform with frequency pre-warping. The low pass filter just after the first squaring block is suggested to be a Butterworth filter of $6^{th}$ order, having a 3 dB frequency of 35 Hz

The low pass is designed to eliminate twice the a.c. system fundamental frequency resulting from squaring (with an attenuation greater than 91 dB at 100 Hz) and thus this limitation in bandwidth allows for a further sample rate reduction. The error introduced in reproducing the magnitude response of the analogue weighting filter by the digital filter is also linked to the sampling rate requiring about 200 Hz to provide a signal free from aliasing.

The digital flickermeter offers the flexibility to switch between different weighting filter characteristics. The respective weighting filters for 120V and 230V systems were both implemented as indicated in Figure 1.

The final low pass filter has the characteristic of a $1^{st}$ order RC circuit with a time constant of 300 ms [1]. Its conversion to a discrete-time filter by the bilinear transform is not critical. The corresponding cut-off frequency is 0.53 Hz and the sampling frequency can be lowered to 50 Hz, the minimum rate for IFL samples to be supplied to the statistical evaluation block according to the IEC standard.

The following subsections discuss the chosen sampling frequencies and the digital weighting filter design.

### 2.1 Sampling Frequency

The proposed digital flickermeter employs multirate processing. The use of the lowest possible sample rate at each stage ensures maximum computational efficiency.

Flicker is considered to be an amplitude modulation of the ac system voltage exhibiting the frequency spectrum of a carrier with sidebands. Loads that draw varying currents (e.g. arc furnaces) produce this type of disturbance. Under this assumption and for a maximum flicker fusion frequency of 35 Hz, the interesting frequency band for flicker measurements extends up to 85 Hz for a 50 Hz ac system (95 Hz for a 60 Hz ac system). The input bandwidth of the flickermeter should however not be restricted to this small range because there are other phenomena that have to be taken into account.

Interharmonics also produce light flicker [9]. Beating of certain components present in the frequency spectrum of the voltage can lead to voltage amplitude fluctuations causing perceptible light flicker. For example a voltage with a spectrum consisting of the ac system fundamental, a 5th harmonic, and a 260 Hz component will contain a 10 Hz beat frequency in its envelope relevant for flicker perception, although the spectrum has no components at frequencies below or close to the fundamental.

Spectral components at non-harmonic frequencies can result from harmonic cross-modulation in the process of ac-dc static conversion. In practice the amplitudes are very small (below 1 %) so that a single interharmonic is unlikely to cause perceptible flicker. The interesting frequency range is therefore limited to approximately 20-80 Hz if only the beating of single interharmonics with the fundamental is considered [9]. However several interharmonics and harmonics might result in noticeable flicker because the individual flicker levels are additive. Consequently flickermeters should have a sufficiently high input bandwidth.

The bandwidth of our digital flickermeter is 1.2 kHz limited by a $255^{th}$ order linear phase FIR anti-aliasing filter designed by the Remez exchange algorithm. The stopband extends from 1.6 kHz allowing for a sample rate reduction to 3.2 kHz achieved by a decimation factor of~16 from the sample rate 51.2 kHz at which the analogue-to-digital converter of CHART operates [7]. The FIR filtering was embedded in the down sampler and filter coefficient symmetry was exploited to achieve an efficient decimator.

The first LP filter of the IEC flickermeter has an attenuation of at least 90 dB at 100 Hz and above to attenuate twice the fundamental frequency component resulting from squaring. Therefore the sample rate can be reduced further after the filter. The remaining filters operate at a sample rate of 200 Hz. The final LP filter reduces the signal bandwidth again (1st order filter, $f_{cut\text{-}off}$=0.53Hz) so that the sample rate can be lowered to minimise computations involved in the statistical

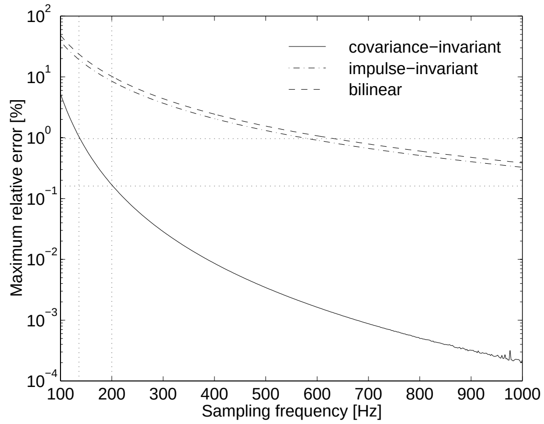

Figure 2: Magnitude errors of digital filter designs compared to the analogue filter vs. sampling frequency

evaluation of the IFL. The minimum rate at which samples should be supplied to the statistical evaluation block is 50 Hz according to [1].

### 2.2 Digital Weighting Filter Design

The magnitude response of the digital weighting filter influences the overall characteristic of the meter strongly and therefore the digital filter should match the IEC specification closely.

The bilinear, impulse-invariant, and covariance-invariant transforms were considered for the analogue to digital filter conversion. Figure 2 shows the maximum relative errors in magnitude for digital weighting filter designs referenced to the IEC analogue filter specification. The filters were designed for a range of sample rates and the maximum occurring error in the interesting frequency range (0.5-35 Hz) was recorded each time. The superior magnitude response fit of the covariance-invariant filter can be confirmed [10]. For the chosen sample rate of 200 Hz only the covariance-invariant filter gives an acceptable accuracy.

The dc rejection of the digital filter given directly by the covariant-invariant transform has been improved by shifting the relevant filter zero onto the unit circle (i.e. from 0.9989+j0 to 1.0+j0). This eliminates the need for a separate HP filter preceding the weighting filter. The determined filter coefficients are given in Table 1.

As already mentioned, the original filter had been specified for 230 V lamps preventing the use of the flickermeter in 120 V systems. A digital flickermeter has the flexibility to switch between different weighting filter characteristics indicated in Figure 1. Table 2 gives the coefficients for the digital filter corresponding to the to 120 V systems adapted weighting filter of [11]. Measurements using the appropriate filter will be consistent and allow for the comparison of results.

Table 1: Coefficients of weighting filter for 230 V systems (sample rate: 200 Hz)

| Order | Numerator Coefficient | Denominator Coefficient |
|---|---|---|
| $z^0$ | +9.487215e-02 | +1.00000 |
| $z^{-1}$ | -1.582865e-01 | -3.167151e+00 |
| $z^{-2}$ | +4.023729e-02 | +3.752054e+00 |
| $z$-$^3$ | +2.317702e-02 | -1.958255e+00 |
| $z^{-4}$ | | +3.747149e-01 |

Table 2: Coefficients of weighting filter for 120 V systems (sample rate: 200 Hz)

| Order | Numerator Coefficient | Denominator Coefficient |
|---|---|---|
| $z^0$ | +6.568031e-0 | +1.00000 |
| $z^{-1}$ | -1.081596e-01 | -3.236145e+00 |
| $z^{-2}$ | +2.660934e-02 | +3.936870e+00 |
| $z$-$^3$ | +1.586997e-02 | -2.127023e+00 |
| $z^{-4}$ | | +4.275761e-01 |

### 2.3 Compliance Test

To verify the compliance of the developed flickermeter with the IEC standard, the standardised tests were conducted using a MATLAB simulated flickermeter and the CHART implementation.

Figure 3 represents (in solid line) the maximum IFL for sinusoidal flicker over the range of frequencies and magnitudes specified by the IEC. All values lie well within the ±10 % tolerance band around one unit of perceptibility. This is equivalent to the criterion of the IEC standard which demands that for unity output each test signal amplitude must lie within ±5 % of the reference value.

If CHART's analogue input circuitry is bypassed for

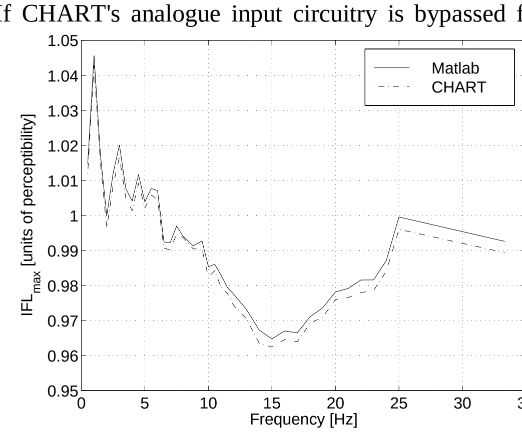

Figure 3: Compliance test results for the simulated flickermeter, MATLAB and the CHART implementation. Sinusoidal test signal.

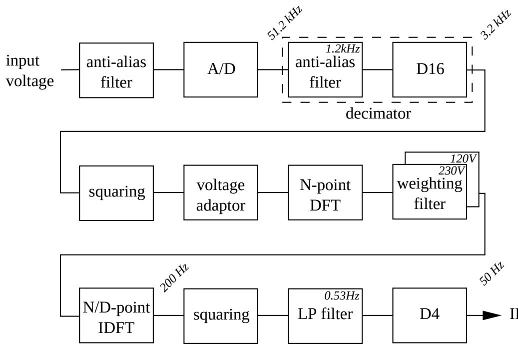


Figure 4: Block diagram of the digital frequency domain based flickermeter (statistical evaluation excluded). Sampling frequencies are indicated in italics.

the tests, thus supplying the digital test signals directly to the DSPs, the test results coincide with the ones obtained from the MATLAB simulations in Figure 3. This is hardly surprising since both systems are fully digital differing only in their floating point number representation (32 compared to 64 bits).

The addition of a custom built a digital signal generator including a 16-bit digital-to-analog converter with a resolution of about 30 ppm provided the test results shown in Figure 3 (dotted line), clearly they match those of the MATLAB simulation results well and lie well within the permitted tolerance.

## 3. DIGITAL F-DOMAIN BASED FLICKER-METER

### 3.1 Review of Existing Proposal

Frequency domain based flicker measurement has previously been found computationally more efficient and capable of replicating the analogue weighting filter response exactly [6].

The method uses a pruned Fast Fourier Transform (FFT) to estimate the spectrum of the modulating flicker, i.e. the sideband of the amplitude modulated ac system fundamental. Subsequently the weight factors and squaring are applied to each spectral component. The sum of all components gives an estimate of the average IFL. The FFTs are performed on data accumulated over 2.13 s.

This approach fails to provide the time domain IFL needed for the subsequent statistical evaluation. The very restricted input bandwidth of 30-60 Hz (for a 60 Hz ac system) accounts for flicker due to amplitude modulation only and strict compliance with the IEC standard is not achieved. Moreover the long time windows used inherently assume a constant disturbance, which is not realistic.

### 3.2 New Proposal

In comparison to the time-domain flickermeter presented in the previous section (Figure 1), Figure 4 shows that the weighting filter, preceded by a Discrete Fourier Transform and followed by its inverse, has been realised by the f-domain filtering approach. The demodulation low pass filter has been made redundant.

The implementation via the DFT will generally increase the efficiency for FIR (Finite Impulse Response) filters, especially for high filter orders. The anti-aliasing filter which is part of the decimator is thus also a candidate for a f-domain realisation. However in this case the gains in efficiency achieved by the combination of the filter with the down-sampler would be lost.

The realisation of the squaring block(s) in the f-domain is possible but also very inefficient. Because multiplicative modifications correspond to convolution in the opposite domain, the squaring would involve a filtering operation on the spectrum. This would degrade the efficiency of the system to a higher degree than the gains achieved in other parts by the use of DFT filtering.

The spectrum produced by the N-point DFT just before the weighting block, consists of N evenly spaced frequency samples covering the range from d.c. up to the Nyquist frequency of the signal, which equals half the sampling frequency i.e. 1.6 kHz. Recalling, that the weighting filter combined with the original low pass filter has been designed to attenuate frequencies equal to and greater than twice the a.c. system fundamental frequency which are not perceptible, this is clearly unnecessary and therefore computationally inefficient. Instead, a DFT with a reduced number of outputs is employed extracting only the lower end of the spectrum, namely the range from 0-100 Hz. Specialised FFT algorithms such as pruning, transform decomposition [12]..etc,. are available to perform this task.

At this point it is also important to realise that the DFT returns negative frequencies, which for real input are redundant. Real-valued FFTs exploit this redundancy, to once again increase computational efficiency.

The IDFTs are performed on the reduced spectrum (0--100 Hz) resulting in a sampling rate reduction by a factor of 16. The remaining blocks thus operate at 200 Hz as for the time domain realisation. In the frequency domain the weighting filter is simply realised by multiplicative changes to the computed spectrum. Thus the number of (complex) weight factors equals the number of frequency samples that represent the spectrum produced by the DFT. The frequency samples are spaced $\Delta f$ apart, where

$$\Delta f = \frac{f_s}{N} \qquad (1)$$

For example for $\Delta f = 1$ Hz, there will be 101 frequency samples in the 0-100 Hz range, so that the weights only have to be calculated once from the analogue weighting filter response and can then be stored in a look-up table.

The frequency resolution improves with increased N (Equation 1) and thus spectral leakage will decrease. At the same time the ability to resolve changes over time in the analysed signal will diminish. Since flicker represents disturbances of stationary as well as non-stationary nature, sufficient temporal resolution is, however, required. Therefore the choice of N has to be a compromise. It has to be noted, that the frequency resolution of the DFT cannot be increased by simply zero padding the input sequence to length N because it does not increase the time duration of the signal. This follows from the uncertainty principle which states that the uncertainty in measuring a frequency is the reciprocal of the time taken to measure it [13]. Zero padding only alters the sampling instants $k.\Delta f, (k = 1,...N)$ of the discrete spectrum.

N also determines the amount of time domain aliasing in the process of reconstructing the time domain waveform via the IDFT. This effect is the analogue to the more familiar frequency domain aliasing, which occurs when a signal is sampled with a frequency lower than twice its maximum frequency (bandwidth). Similarly the spectrum has to be sampled (in frequency) at a rate of at least twice the signal's ``time width'' to prevent aliasing. Another constraint is that N has to be a power of two to allow for the computation of the DFT by efficient FFT algorithms.

### 3.3 Compliance Tests

The f-domain flickermeter was simulated using MATLAB running under UNIX. An implementation on CHART's DSPs was not undertaken, since both systems are fully digital and only very small differences due to the single precision number representation on the DSP are expected.

Figure 5 presents the test results using 8192-point FFTs, and a window length $N_h$ of 4096. To take into account the proposed inclusion of a test signal at 33.33 Hz to achieve better consistency between different flickermeter implementations the response of the weighting and the demodulation low pass filter in the DFT filter bank are combined.

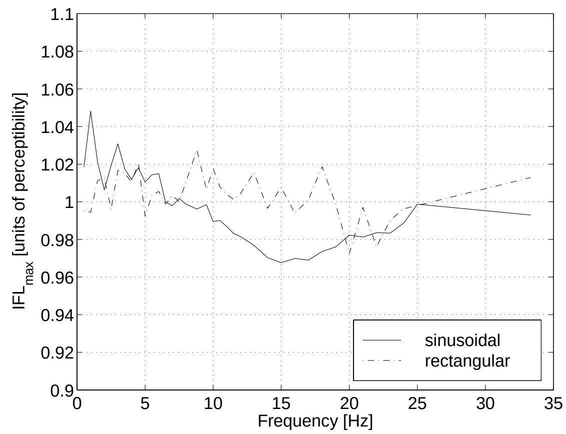

Figure 5: Compliance test results for the f-domain flickermeter. Hamming analysis windows and rectangular synthesis window used. $N=N_f=8192$.

Compliance with the IEC standard, which currently only comprises the test signals in the 0-25 Hz frequency range, is proven for those configurations, since the curves lie within the tolerance band extending from 0.9-1.1 units of perceptibility.

## 4. PRACTICAL APPLICATION

To test the flickermeter in a more realistic environment, arc-furnace voltage recordings were used as input Arc-furnace flicker has varying spectral content which is of particular interest when testing an FFT based application, otherwise problems due to spectral leakage may not be recognised.

The IFL signals obtained with the time domain and frequency domain implementations were compared using the IAE as a figure of merit. Figure 6 represents the resulting output signals and their relative differences and that there is no appreciable difference between the two IFL curves. The errors are small reaching only a few percent. This manifests itself in the IAEs which evaluate to 0.078 % and 2.97 % respectively.

Note that the overlap used in Figure 6 is only 50 % and thus lower than the theoretical value. The results for an overlap of 75 % show no appreciable difference.

## 5. CONCLUSION

This paper has presented digital flickermeter designs compliant with the IEC standard that use either the time or frequency domain. Both approaches model the lamp-eye-brain chain ensuring realistic flicker measurements for any kind of disturbance. Maximum computational efficiency is achieved by decimating to the lowest appropriate sample rate at each stage.

A complete implementation of the flickermeter in the f-domain is not practical due to the high computational requirements involved. However, the implementation of the weighting filter via the DFT provides flexibility as any desired filter response can be quickly accommodated.

The flexibility of a digital flickermeter allows changes to the weighting filter characteristic according to different light bulb parameters, and it can therefore be used in 230 V and 120 V systems. The digital weighting filter for the time domain realisation was obtained from the analogue prototype by the covariant-invariant transform, which provides a high magnitude response accuracy. The frequency domain based meter realises the weighting filter response exactly and, moreover, any arbitrarily shaped response can be easily adopted.

The consistency of the measured IFL with the time and frequency domain implementations has been verified using sinusoidal and arc-furnace flicker. The latter provides a realistic test signal with variable spectral content necessary to verify sufficient time resolution of the system.

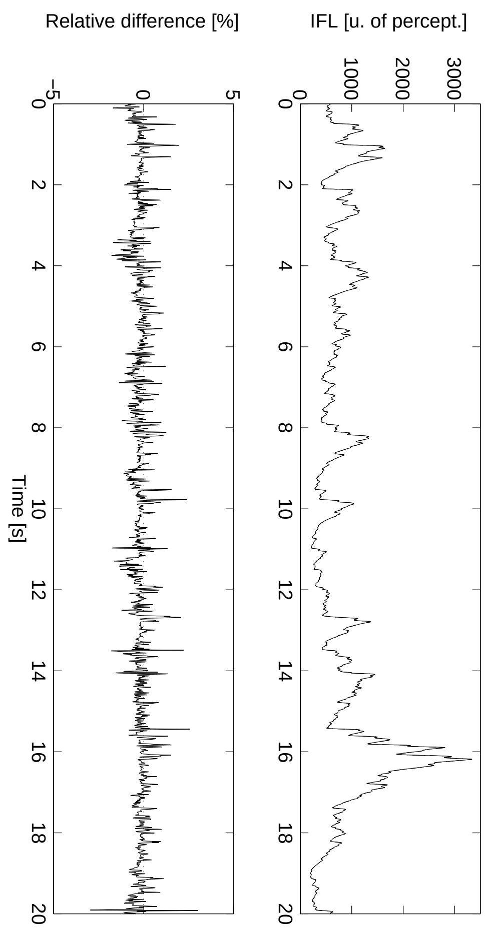


Figure 6 IFL resulting from arc-furnace flicker. The solid lines show the output from the time domain flickermeter, the dash-dotted ones indicate the IFL from the f-domain approach. IAE=0.078,$N_h$=4096,L=2048 (50% overlap)

## 5. REFERENCES


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