Filterbank Multicarrier Refletometry for Cognitive Live Wire Testing

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Abstract

Multicarrier reflectometry (MCR) for locating faults on live wires has recently been proposed. This paper studies the use of filterbanks for generation/synthesis of MCR test signals and also for signal analysis for fault identification and location. We note that the test signals have to be confined to the portion(s) of the frequency band that is (are) free of signals already on the wires in order to avoid interfering with them. Moreover, for effective analysis of the reflected waves, optimal filters that separate the test signal tones and also minimize leakage from the existing signals on the wires should be designed. We discuss the criteria necessary to design effective MCR systems and develop the relevant filterbank design procedures. We also introduce the novel idea of cognitive live wire testing, where the tester first measures the live wire signal activities and then decides on which part of the spectrum should be used for testing.

I. Introduction

Aging wiring is one of the issues in the forefront of maintaining older aircraft. Incidents such as Swiss Air Flight 111 and TWA Flight 800 have led to a heightened public awareness of these critical wiring issues [1]. Conventional wire inspection is typically limited to visual inspection during routine maintenance or highly invasive testing during major modifications. But visual inspection misses many flaws, especially within wire bundles. Therefore, methods to locate small faults before they create system level problems are desired.

A new class of reflectometry methods is being developed to locate intermittent fault during flight, where they most often occur, [2]–[6]. These systems must be fast enough to locate faults lasting only a few milliseconds, must be robust enough to be impaired by the existing signals on the wires or the inherent electromagnetic noise on aircraft, and must also be compatible with the existing aircraft signals so as not to interfere with them. Among the new live wirw reflectometry methods is multicarrier reflectometry (MCR), [2].

Reflectometry is essentially a system identification problem. A test signal is sent down the wire under test, and the reflected signal is used to study the condition of the wire, [2]–[6]. The wire may be treated as a plant whose input is the test signal, and its output is the reflected signal. Identification of the plant, or equivalently, the condition of the wire, is the goal. The plant response can be obtained by comparing its input and output signals. In time domain reflectometry (TDR), for example, the input is chosen to be a step function or impulse, and accordingly the plant output/reflected signal is the step or impulse response of the plant/wire [5]. In MCR, the test signal is chosen such that the plant response can be measured in the frequency domain. The test signal is the summation of a number of tones, and the plant/wire frequency response is measured by comparing the amplitude and phase of the transmitted tones with their reflected counterparts [7]. This method of system identification was applied to wire testing in [2]. An MCR test equipment selects one or more portions of the spectrum that is/are free of live wire signals for transmission of the tones/test signal. This leads to frequency agility and adaptability – ideal for live wire testing.

The work presented in [2], although pioneering in the sense that it first introduced the concept of MCR, has not given due attention to the optimality of the test signals and the filters to be used for signal analysis. This paper concentrates on these aspects of MCR and refines the MCR method that was proposed in [2].

The paper is organized as follows. In Section II, we describe the problems of out-of-band interference in signal synthesis (i.e., interference with the live wire signals) and in-band interference while performing signal analysis (i.e., the interference that tester receives from the live wire signals). The proposed filterbank solution and the filter design methodologies are presented in Section III. The concept of cognitive live wire testing is described in Section IV. The concluding remarks are drawn in Section V.

In this paper, the following notations are used. A function x of continuous time t is written as $x(t)$. When $x(t)$ is sampled and thus time is discrete, the resulting sequence is represented as $x[n]$. Vectors are always in column form and are denoted by lowercase bold letters. Uppercase letters are reserved for matrices. Scalar variables are denoted by lowercase non-bold letters. Constant scalars, such as dimensions of a matrix, are denoted by uppercase non-bold letters. Linear convolution is denoted by '$*$'.

II. In-Band and Out-of-Band Interference

In [2], the multicarrier test signal is synthesized by summing a set of sine waves over a finite window of time. Mathematically, this is written as

$$x(t) = \left(\sum_{i \in \mathcal{S}} a_i \sin\left(\omega_i t + \theta_i\right) \right) w(t) \tag{1}$$

where $\mathcal{S}$ is a set of frequency indices, a_i, ω_i and θ_i are the magnitude, frequency and phase of the ith sine wave, t denotes continuous time, and $w(t)$ is a window function that limits $x(t)$ to a finite duration. In [2], $w(t)$ was chosen as a rectangular window with duration of T:

$$w(t) = \begin{cases} 1, & 0 \leq t < T \\ 0, & \text{otherwise.} \end{cases} \tag{2}$$

Moreover, it was suggested that the phase angles θ_i should be chosen to minimize the peak-to-average power ratio (PAPR) of $x(t)$.

Defining

$$v(t) = \sum_{i \in \mathcal{S}} a_i \sin\left(\omega_i t + \theta_i\right) \tag{3}$$

and taking the Fourier transform of $x(t)$, we obtain

$$X(f) = V(f) * W(f) \tag{4}$$

where $*$ denotes convolution.

Using (4) and noting that the Fourier transform of the rectangular window $w(t)$ is a sinc pulse $\left(\mathrm{sinc}(x) = \frac{\sin(\pi x)}{\pi x}\right)$, and the Fourier transform of each sine wave is the sum of two impulses, at frequencies $f = \pm\frac{\omega_i}{2\pi}$, one finds that the energy spectrum of $x(t)$, defined as $|X(f)|^2$, is the sum of a number of sinc-squared pulses. An example of such a spectrum is presented in Fig. 1. Here, the length of the test signal is 320 samples and a multicarrier system with 128 subcarrier is assumed. Subcarrier numbers 8 to 25 and 33 to 45 are used for transmitting the test signal. A relevant characteristic of this spectrum to this paper is the fact that $|X(f)|^2$ is not confined to the intended frequency band. There is a significant level of spectral leakage to the portions of the band that may be occupied by live wire signals; for example, if the live wire signal was in the (normalized) frequency range 0.2 to 0.25 or near DC. In the latter cases, the attenuation of the test signal over the live wire bands is only 20 to 25 dB. Hence, without other measures, the MCR test signal may interfere with the live wire signal. We refer to the interference generated by the test signal on the live wire signal as the out-of-band interference.

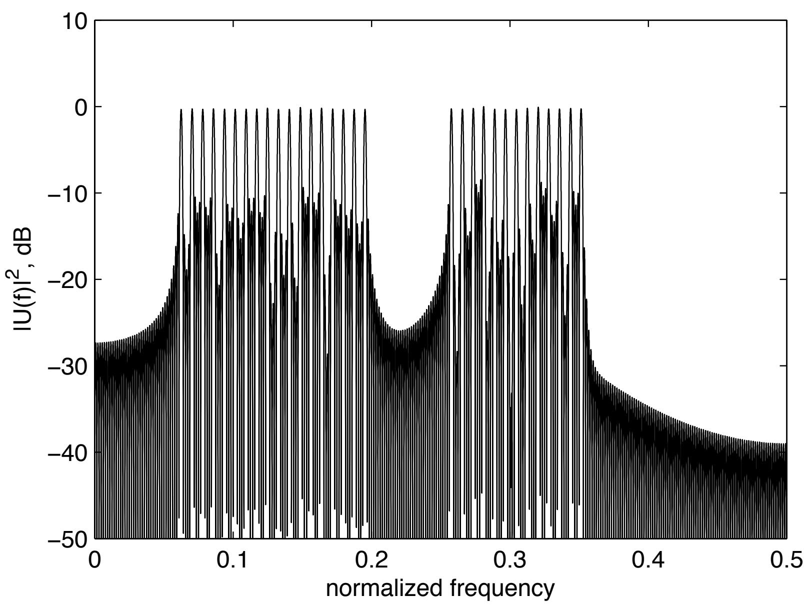

Fig. 1. An example of the MCR test signal from [2].

Another problem will arise in the analysis of signals. In [2], it is proposed that a rectangular windowed portion of the reflected signal may be passed to an FFT processor for analysis. Here again, the application of the rectangular window is equivalent to applying a set of narrow band filters whose frequency responses are sinc pulses. Such responses suffer from poor stop band rejection which in turn means the live wire signal energy is picked up by the analysis filters, reducing the accuracy of the measurements. We refer to this type of interference as the in-band interference.

The problems associated with the rectangular window are well know to the signal processing community, and methods to alleviate them have been studied in different contexts. In spectrum analysis the problem of spectral leakage is solved by replacing the rectangular window with other window types, e.g., a Hamming or Hanning window, [8]. For digital subscriber lines

(DSL) some of the problems are very similar to those encountered in live wire testing. There, a communication signal is transmitted over an unshielded telephone line that is exposed to radio signals over some portions of the frequency band. Therefore, similar to our case here, portions of the band should not be used for communication and the synthesized signals should be designed to avoid these radio bands [9]. In the analysis part of the system (the receiver), proper filtering should be performed to eliminate as much of the radio signal - indeed, a very similar problem to the live wire testing. In DSL, the problem of leakage and thus filtering is solved by replacing the rectangular window with a window that tapers at the sides, [10]. This avoids the sharp edges that are the main source of spectral leakage/poor filtering.

Multicarrier modulations have been widely used in wireless communication systems. Of the various multicarrier modulation methods, orthogonal frequency division multiplexing (OFDM), is the most widely adopted in the current standards [11]. Filterbank multicarrier (FBMC) methods are an alternative modulation technique which can be used in wireless communication systems. In cognitive radios, secondary users are required to find spectrum holes to avoid interference with the primary users, also a very similar scenario to live wire testing. It has been shown that FBMC techniques perform better than OFDM in cognitive radios [12], [13]. On the same basis, the work presented in this paper is an improvement over [2].

In this paper, we use the windowing method proposed in [10] to reduce the out-of-band interference. However, for signal analysis we resort to a direct filterbank solution. For live wire signals, the filterbank analysis method is a good choice since we can design (as shown in Section III-B) optimal filterbanks that minimize the in-band interference and assure near-perfect separation of the tones in a test signal. This application of filterbanks is unique to live wire testing, because the test signals are a sum of a set of pure tones. We also study some details of the windowing solution of [10] and develop new insights that lead to a method of improving the window function that has been commonly used in the past. We show that the proposed signal synthesis with windowing may be formulated in the frame work of filterbanks.

III. Filterbank MCR

Fig. 2 presents the various blocks involved in the construction of a filterbank MCR (FBMCR). The output of the signal generator block, $x(t)$, is a multitone finite duration signal whose spectrum is confined within a desired band. The system, in the context of this paper, is a live wire carrying existing signals. The input to the system is the point where the test signal is injected into the wire, and its output is the point where the reflected signal from the wire is seen; for details of the relevant circuit diagrams see [2]. The live signal on the wire is also seen at the output of the system. This is indicated in Fig. 2 as an additive interference, considering the fact that the live signal is a source of interference to the intended measurement/FBMCR. The analysis filterbanks (AFBs) extract the signal tones applied at the input as well as their replicas after reflecting from the end if the wire or its fault, i.e., at the system output, $y(t)$. Comparing each pair of outputs from the AFBs, as discussed in [2] (also, see [7] for more details), one can estimate the samples of the frequency response of the system (the wire). The measured frequency response is then used to study the time domain characteristics of the wire. These operations are performed in the system identification block. Details of how this conversion can be performed are given in [2]. In the following, we present the details of the signal generator and ABF blocks of Fig. 2 and discuss the criteria that should be used to design these blocks optimally.

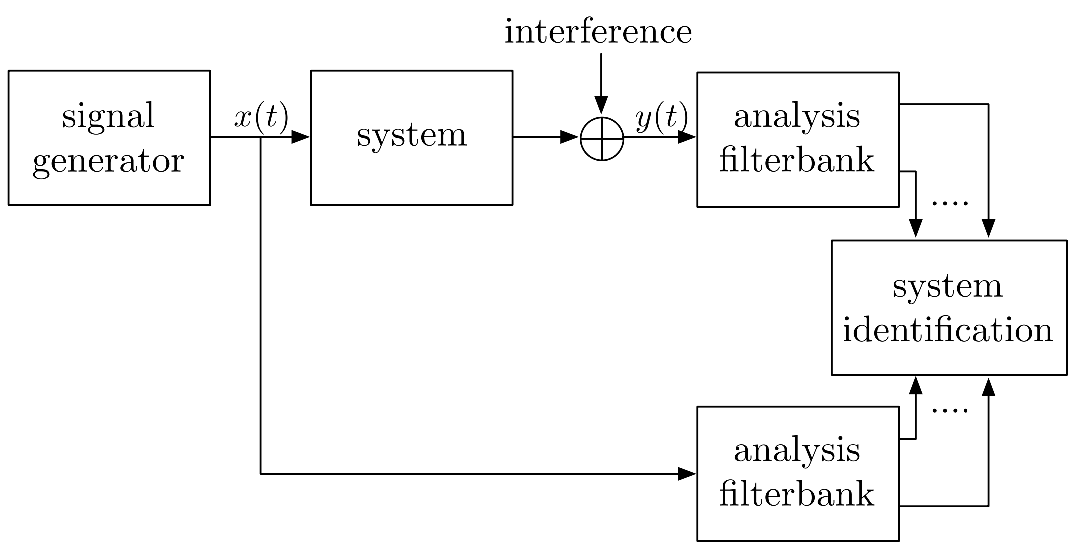

Fig. 2. General structure of a filterbank MCR system

A. Signal generator

The test signal $x(t)$ is synthesized by summing a set of tones that are spread over a selected frequency band. $x(t)$ should have the following properties:

1) It should have a finite duration.
2) Its spectral content should be confined within the selected band, minimizing its out-of-band interference.
3) It should have a minimum peak-to-average power ratio (PAPR).

The first property is required as any realistic test has to be completed within a finite time. This finite time can be easily set by truncating $x(t)$ (or, equivalently, the tones) outside of a preselected interval.

Once the tones are truncated, their spectrum will spread across the frequency axis, (according to (4)), and as a result the spectrum spreads over the frequencies that are not part of the intended band. We minimize the out-of-band interference by first adding the tones and then applying a well-designed window function to the summation.

The PAPR is minimized through a three step signal synthesis. We first choose a discrete time periodic signal $\phi[n]$ with a small PAPR and with a period of N samples. Such a periodic signal has a Fourier series with N harmonics at frequencies $\frac{2\pi k}{N}$ for $k = 0, 1, \cdots, N-1$. It is also desirable to choose $\phi[n]$ such that its harmonics have equal power. This results in equal excitation at all frequencies where we wish to estimate the frequency response of the system. Equal excitation is desirable and commonly used in system identification, as it results in equal accuracy at all the test points [14]. Fortunately, sequences with small PAPR and equal power harmonics exist. One class of such sequences are the so called *polyphase codes* devised by Chu [15]. They are defined as

$$\phi[n] = \begin{cases} e^{i\frac{K\pi n^2}{N}}, & N \quad \text{even} \\ e^{i\frac{K\pi n(n+1)}{N}}, & N \quad \text{odd} \end{cases} \tag{5}$$

where K is a number relatively prime to N. Note that $\phi[n]$has the instantaneous power of unity for any value of n. Hence, its average power is also unity, which gives a PAPR of 1 (0 dB). This, clearly, is the lowest PAPR possible.

In the second step, we find the Fourier series coefficients of $\phi[n]$ by taking the DFT of one cycle of it. The result which we call $\Phi[k]$ is used to generate the continuous time version of $\phi[n]$ as

$$\phi(t) = \sum_{k=0}^{N-1} \Phi[k] e^{j\frac{2\pi k}{N}t}. \tag{6}$$

This is a baseband complex-valued signal that may then be up-converted to any arbitrary band according to the equation

$$v(t) = \Re\{\phi(t)e^{j2\pi f_c t}\} \tag{7}$$

where f_c is the carrier frequency (the center of the band) and $\Re\{\cdot\}$ denotes the real part of.

We note that after interpolation, the PAPR is no longer at the minimum level 0 dB. The interpolated signal always has a PAPR which is greater than 0 dB. However, we argue that the use of polyphase codes, which start with a minimum PAPR, leads to a continuous time test signal with a small PAPR. We also note that different polyphase codes result in different PAPR values and thus we examine all choices of $\phi[n]$ for each code length, N, to find the one that results in minimal PAPR. The test signals generated in this way (we examined values of N in the range of 10 to 100) have PAPR values of around 2.5 dB, which is relatively small. To compare, the phase randomization method used in [2] results in signals with PAPRs that are 1 to 2 dB larger.

Recalling that the test signal should have a finite duration, in the third step of test signal generation, a time window should be applied to $v(t)$, as in (1). The common window suggested in the literature is the raised cosine function defined as

$$w(t) = \begin{cases} \dfrac{1-\cos(\frac{\pi t}{T_0})}{2}, & 0 \le t \le T_0 \\ 1, & T_0 \le t \le T \\ \dfrac{1+\cos(\frac{\pi(t-T)}{T_0})}{2}, & T \le t \le T+T_0 \\ 0, & \text{otherwise} \end{cases} \tag{8}$$

where T_0 is the width of roll-off time at each side of the window, and $T+T_0$ is the total duration of the window. Although the raised cosine window is commonly used in signal processing, it is not necessarily the best window function. An analysis of $w(t)$ leads to some insight that we use to propose an improved window function.

1) An analysis of the raised cosine window function: When the window function $w(t)$ is chosen according to (8), one finds that

$$W(f) = T\text{sinc}(fT) \times \frac{\cos(\pi f T_0)}{1-4f^2T_0^2} \times e^{-j\pi f(T+T_0)}. \tag{9}$$

This is obtained by noting that

$$w(t) = \Pi\left(\frac{t-T/2}{T}\right) * h(t) \tag{10}$$

where

$$\Pi\left(\frac{t}{T}\right) = \begin{cases} 1, & |t| \leq T/2 \\ 0, & \text{otherwise} \end{cases} \tag{11}$$

and

$$h(t) = \frac{\pi}{2T_0} \sin\left(\frac{\pi t}{T_0}\right) \Pi\left(\frac{t - T_0/2}{T_0}\right). \tag{12}$$

Clearly, the first term on the right-hand side of (9) is the Fourier transform of $\Pi(t/T)$. The second term is obtained by taking the Fourier transform of $h(t + T_0/2)$. The last term (a linear phase term) arises because the two terms on the right-hand side of (10) are the time shifted versions of $\Pi(t/T)$ and $h(t + T_0/2)$, respectively.

It is instructive to note that the Fourier transform of $h(t)$ appears as a multiplicative factor to the sinc pulse on the right-hand side of (9) and thus has the role of attenuating the side lobes of $\text{sinc}(fT)$. A good choice of $h(t)$ can significantly improve the out-of-band interference of the spectral content of $x(t)$.

To develop an in-depth understanding of the impact of $h(t)$ on the magnitude response of $W(f)$, in Fig. 3 we have presented $|W(f)|$, $|T\text{sinc}(fT)|$ and $|H(f)|$. The relationship between the parameters T and T_0 and some of the features of these plots are also presented. In particular, we may note that the width of the main lobe of $|H(f)|$ is equal to $1.5/T_0$ and, thus, decreases as T_0 increases. On the other hand, narrowing this width is desirable, as it results in attenuating more of the side lobes of $|T\text{sinc}(fT)|$, thus, improving the out-of-band interference of the spectrum of $x(t)$.

2) An optimal choice of the window function: Looking at $h(t)$ in the time domain gives us a different prospect. $h(t)$ is a half sine wave with the sharp edges at $t = 0$ and T_0. Such sharp edges constitute high frequency components in its Fourier transform, resulting in a magnitude response $|H(f)|$ with side lobes that may not be sufficiently small. Hence, the function $h(t)$ as given in (12) may not be a good choice and, thus, a more elegant design of it can lead to better spectral containment of $x(t)$ within the desired band. We argue that an optimal choice of $h(t)$ can be obtained by optimizing $h(t)$ such that the total energy of the side lobes of $H(f)$ is minimized. This leads to a choice of $h(t)$ which is known as a prolate function [16]. Prolate functions are flexible in the sense that for a given time span T_0, the width of the main lobe in the frequency domain can be traded for different suppression level of their side lobes. On the contrary, the choice of $h(t)$ in the raised cosine window, for a given time span T_0, results in

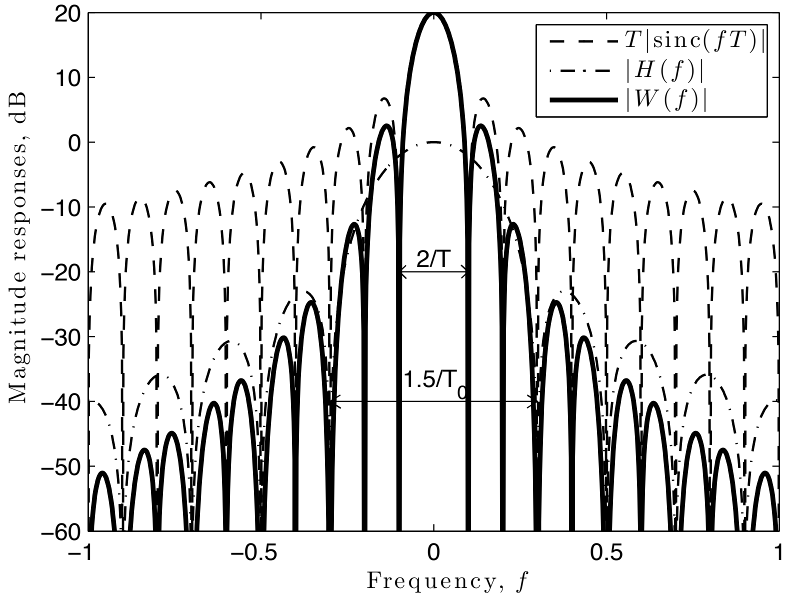

Fig. 3. An example of the magnitude response of the window function $w(t)$ and its factors.

a fixed response in the frequency domain where the width of its main lobe is fixed at $1.5/T_0$. Fig. 4 compares the magnitude responses of $h(t)$ when it is selected to be a prolate function and when it is selected according to (12), both with the same time duration T_0. The prolate design that we have selected has a slightly wider main lobe in order to achieve higher attenuation of its side lobes. Here, at the cost of a slightly wider main lobe, we achieve 10 to 20 dB improvement in suppression of the side lobes.

To conclude our discussion in this section, we present in Fig. 5 a repeat of the spectrum that we previously presented in Fig. 1 after applying a window function based on the proposed prolate design. The length of the window function here is the same as the length of the rectangular window in Fig. 1; 320 taps. As seen, and expected, the out-of-band spectra are significantly attenuated from the values between -20 dB and -40 dB to below -100 dB. As a result the interference from the test signal to the signal on the live wire is orders of magnitude smaller and thus will be negligible.

3) The relationship between time domain windowing and filterbank synthesis: According to the procedure discussed above, the test signal $x(t)$ is obtained by adding a number of tones and then applying a window function to the result. Concentrating on the ith tone in $x(t)$ which we

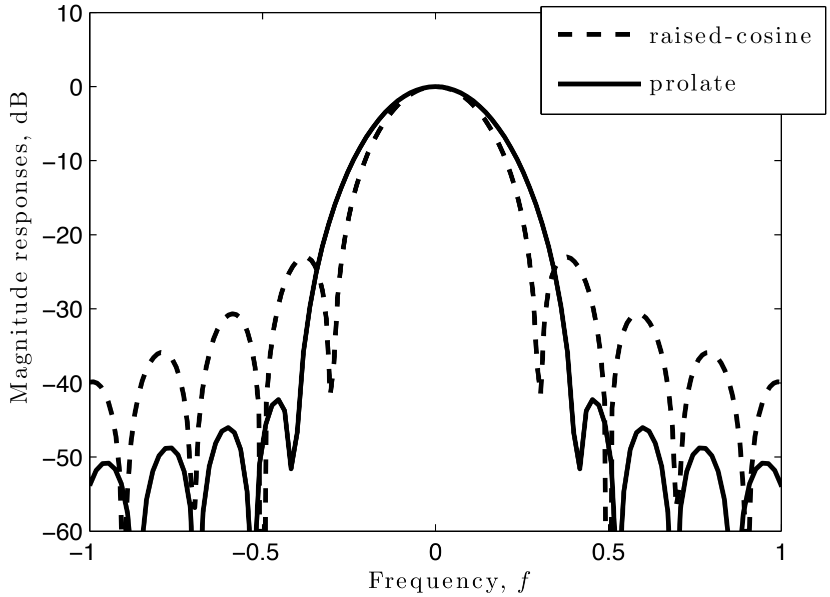

Fig. 4. Magnitude responses of $H(f)$ according to a raised cosine design and prolate design.

call $x_i(t)$, we obtain

$$x_i(t) = w(t)e^{j\omega_i t}. \tag{13}$$

Substituting (10) in (13), we get

$$\begin{aligned} x_i(t) &= \left(\Pi\left(\frac{t - T/2}{T}\right) * h(t)\right) e^{j\omega_i t} \\ &= e^{j\omega_i t} \int_{-\infty}^{\infty} h(\tau)\Pi\left(\frac{t - T/2 - \tau}{T}\right) d\tau \\ &= \int_{-\infty}^{\infty} \left(h(\tau)e^{j\omega_i \tau}\right) \left(\Pi\left(\frac{t - T/2 - \tau}{T}\right) e^{j\omega_i (t-\tau)}\right) d\tau \\ &= \left(h(t)e^{j\omega_i t}\right) * \left(\Pi\left(\frac{t - T/2}{T}\right) e^{j\omega_i t}\right). \end{aligned} \tag{14}$$

This result clearly shows that the ith tone in $x(t)$ is obtained by first applying a rectangular window to the continuous time tone $e^{j\omega_i t}$ and then passing the result through the modulated filter $h(t)e^{j\omega_i t}$, centered at ω_i. Hence, the time domain windowing has an equivalent filterbank synthesis where each tone passes through a filter in the filterbank that is centered at its respective frequency.

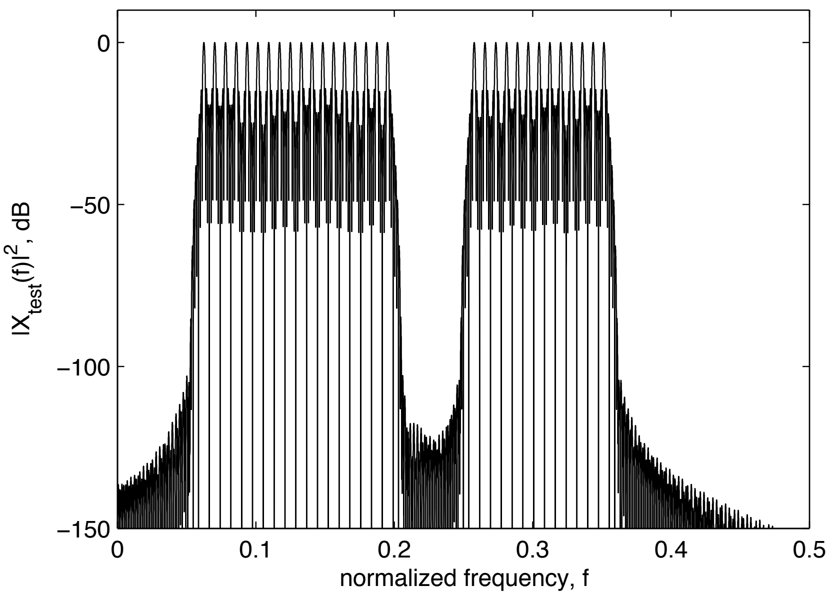

Fig. 5. An example of the test signal after applying $w(t)$. The window function $w(t)$ is based on a prolate design. Compared to Fig. 1, the use of prolate window here has resulted in a significant suppression of the out-of-the-band spectra.

B. Analysis filterbanks

Although the test signal, $x(t)$, and the reflected signal from the wire, $y(t)$, are continuous time, they must be sampled for processing. The sampled signals $x[n]$ and $y[n]$ are analyzed by passing them through a pair of similar filterbanks. The filterbanks are constructed based on a prototype filter $g[n]$. The prototype filter is a lowpass filter with the frequency response $G(f)$, where f is the normalized frequency with respect to the sampling rate. The filterbank is realized by implementing a set of filters with the responses $G(f - f_i)$, $i \in \mathcal{S}$, where f_is are the normalized frequencies of the tones in the test signal. We assume that $x(t)$ and $y(t)$ are demodulated to baseband, i.e., the carrier frequency f_c is removed, before being sampled and passed to the AFBs. Hence, $f_i = \frac{2\pi i}{N}$, for $i \in \mathcal{S}$. This choice allows efficient realization of each AFB in a polyphase structure, [17]. Using polyphase structures, efficient implementation of filterbanks have been applied to different communication and signal processing applications, [17], [18]. The sampling rate at the receiver is an implementation decision which is determined based on system design characteristic such as the analog-to-digital convertor (ADC) sampling rate and power requirement. Critically sampled signals or over sampled signals may be used. It has been shown in [19] that oversampling might result in simplification of the filters used in a multicarrier system.

We propose to design $g[n]$ based on the following criteria:

1) To avoid in-band interference, maximize the stop band attenuation of $G(f)$.
2) Set the gain of $G(f)$ at $f = 0$ (this is at the middle of its pass band) equal to one.
3) The filter frequency response should have zero crossings at integer multiples of the frequency spacing of the tones to avoid interference among different tones. The carrier spacing, here, is the same as carrier spacing of the signal synthesis module.

To cast the above criteria in a mathematical form, we proceed as follows.

1) To maximize the stop band attenuation of $G(f)$, we define the cost function
$$E_s = \int_{f_0}^{1-f_0} |G(f)|^2 df \tag{15}$$
where f_0 is the band edge frequency of the stop band of $g[n]$. Defining the vectors $\mathbf{g} = [g[0]\ g[1]\ \cdots\ g[N-1]]^{\mathrm{T}}$ and $\mathbf{e}(f) = [1\ e^{-j2\pi f}\ \cdots\ e^{-j2\pi(N-1)f}]$ and noting that $G(f) = \mathbf{g}^{\mathrm{T}}\mathbf{e}(f)$, straightforward manipulations leads to
$$E_s = \mathbf{h}^T \mathbf{\Psi} \mathbf{h} \tag{16}$$
where $\mathbf{\Psi}$ is an N-by-N matrix with the klth element of
$$\psi_{kl} = \begin{cases} 1 - 2f_0 & k = l \\ -2f_0 \mathrm{sinc}\left(2f_0\left(k-l\right)\right) & k \neq l. \end{cases} \tag{17}$$
2) To set the gain of $G(f)$ at $f = 0$ equal to 1, the following constraint should be imposed
$$\mathbf{c}_0^{\mathrm{T}}\mathbf{h} = 1 \tag{18}$$
where $\mathbf{c}_0$ is a column vector of length N and elements of 1.
3) The constraints that imposes zero gain at integer multiples of the frequency spacing between the tones are obtained as
$$\mathbf{c}_i^{\mathrm{T}}\mathbf{h} = 0, \quad \text{for } i = 1, 2, \cdots, N-1, \tag{19}$$
where $\mathbf{c}_i$ is a column vector with the elements of 1, $e^{-j2\pi f_i}$, $\cdots$, $e^{-j2\pi(N-1)f_i}$.

The above results suggest that to design the prototype filter $g[n]$, one should minimize E_s subject to the constraints (18) and (19). Hence, using the method of Lagrange multipliers, we define

$$\xi = \mathbf{h}^{\mathrm{T}}\mathbf{\Psi}\mathbf{h} + \lambda_0(\mathbf{c}_0^{\mathrm{T}}\mathbf{h} - 1) + \sum_{i=1}^{N-1} \lambda_i \mathbf{c}_i^{\mathrm{T}}\mathbf{h} \tag{20}$$

and to find $\mathbf{h}$, we form and solve the following set of equations:

$$\nabla_{\mathbf{h}}\xi = 0 \quad \text{and} \quad \frac{\partial \xi}{\partial \lambda_i} = 0, \text{ for } i = 0, 1, \cdots, N-1. \tag{21}$$

Fig. 6 presents the magnitude response of a typical prototype filter that is designed using the above procedure. Here, $N = 16$ and the filter length is $L = 8N + 1 = 129$. By increasing the filter length (L), one can arbitrarily control the stop band attenuation. Therefore, using the filterbank analysis structure, the interference from the signals on the live wire to the test signal is not a limiting factor in decreasing the effective signal to interference plus noise ratio (SINR) in our diagnosis system.

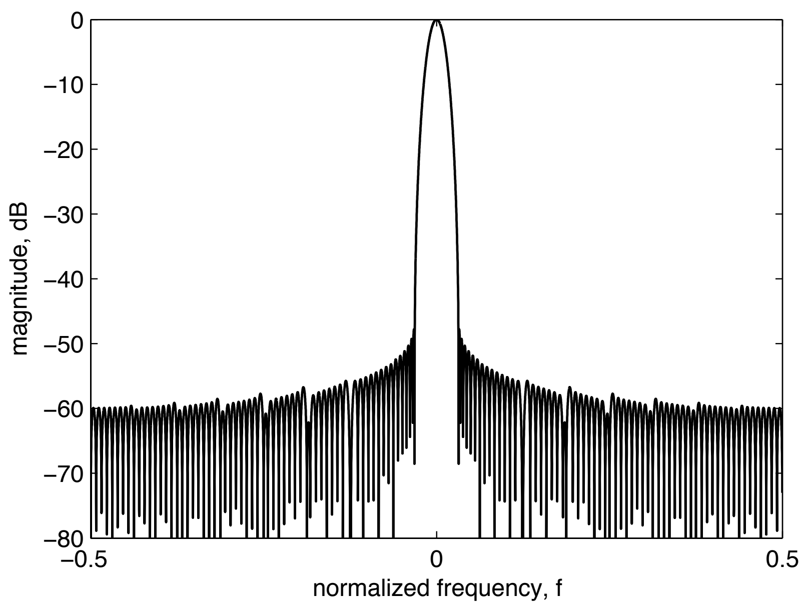

Fig. 6. Magnitude response of the analysis prototype filter.

IV. Cognitive live wire testing

The frequency band(s) over which live wire signals are located in different wires are not necessarily the same and, thus, such bands may be unknown to a live wire tester. We propose to make the tester intelligent by designing it such that before every measurement it tests the spectral activity on the wire and finds a band (or multiple bands) to accommodate the test signal. We borrow this idea from the field of cognitive radio, [12], and thus call the proposed method cognitive live wire testing. Cognitive radios belong to an emerging class of radios that measure the spectral activities in their surrounding environment and begin a communication session over the portion(s) of the spectrum that is not used by other radios. Clearly, extension of this concept to live wire testing is straightforward and needs no further elaboration. We only need to equip the live wire tester with a spectrum sensor, i.e., an spectrum analyzer.

Many spectrum analyzers essentially use a filterbank to extract the spectral energy of the signal at each portion of the frequency band. Moreover, recent studies have identified filterbanks as the natural choice for spectrum sensing in cognitive radios, [18]. We also propose using filterbank method for spectrum sensing in live wire testing, and note that since in our system setup an analysis filterbank is connected to the wire this addition comes at virtually no additional cost.

V. Conclusions

Multicarrier reflectometry (MCR) has recently been proposed as an effective method of testing live wires, [2]. This paper presented a more detailed study of the signal processing tools necessary in the implementation of MCR and discussed how these tools may be perfected. We noted that to minimize the interference between the test equipment and live wire signals, filterbanks are the best signal processing tools. We thus developed novel methods for this application using filterbanks. We showed that using a filterbank based on the prolate window design, the interference level from the test signal to the live wire signal can be controlled and decreased significantly. Hence the effective SINR of the live signal will be minimally affected by the test signal. Furthermore, it was shown that the optimum filters can be designed for analysis filterbanks. The stop band attenuation of the analysis filterbanks can also be made arbitrary small by increasing the length of the filters. We also borrowed an idea from the field of cognitive radios and suggested the concept of cognitive live wire testing.

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