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Cyclic Signals and Systems in Power Line Communications

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ABSTRACT This paper provides a comprehensive discussion about the characteristics of power line communication (PLC) channels related to its synchronization with the mains period. This fact constitutes an interesting and distinguishing feature of these kind of channels. Under the mains voltage presence, the behavior of some loads connected to the power network allows representing the transmission medium as a linear periodically time varying (LPTV) system. Some components in the received noise exhibit periodicity as well, in terms of impulsive waveforms synchronized with the mains period and other cyclostationary (CS) signals. This fact has been taken into account in the design of PLC systems, for instance by defining medium access control (MAC) frames synchronized with the mains period and by adapting the modulation to the cyclic signal-to-noise ratio (SNR) to improve the system performance. Hence, in PLC, it is of particular interest to describe the general relations of LPTV systems and CS signals when both have a common period, which is the purpose of this work. It is not aimed at providing new mathematical tools for the analysis of these cyclic signals and systems, which have been extensively developed over the last five decades, but to serve the readers of the PLC community more accessible mathematical relations that can be applied to the practical problems they have to face. A review of recent works dealing with the cyclic properties of PLC channels and transmission techniques aware of these strategies is included as well.

INDEX TERMS Cyclostationary random signals, time-varying systems, LPTV, power line communications.

I. INTRODUCTION

The term *cyclic* used in the title of this work denotes in a synthetic way both:

- non-stationary random signals but with a periodical correlation, which are usually referred to as *cyclostationary* (CS) signals and,
- *linear periodically time-varying* (LPTV) systems.

There exist valuable classical works devoted to CS signals properties [1], [2] and to LPTV systems [3]–[5] and, even more, works that contain a deep analysis of CS signals, LTPV systems and their interplay, from a comprehensive perspective [6]–[10]. There are also many interesting publications that include discussions on cyclic signals and systems with a scope focused on signal processing for communications [11]–[14]. However, these studies may be too extensive and intricate for a reader attracted by this topic but mostly interested in power line communications applications (PLC) [15], in which cyclic signals and systems can be analyzed from a quite simpler perspective. This paper does not intend to rise to the challenge of such excellent literature on the topic, but it has a more modest purpose: to satisfy a trade-off between rigor and conciseness to yield a reasonably formal treatment of cyclic signals and systems biased to PLC.

CS signals are certainly common in many fields of science, especially in those that study natural phenomena in which periodicity and randomness are inherent. Examples of these are, to name just a few, weather and atmospheric related problems, influenced by rotation and translation of Earth, Moon, etc.; human physiological issues, related to the heart beat rate; acoustic signals or vibrations caused by motors and rotational devices; and, obviously, in communication systems, where modulation and synchronization processes yield to periodicity. In PLC, the study of cyclic signals and systems is justified by the omnipresence of mains voltage, with a period of $T_0 = 1/50 = 20$ ms (in most of the world) or $T_0 = 1/60 = 16.67$ ms (in some American and Asian countries), which affect the characteristics of many connected appliances

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and, consequently, to the communication channel behavior. In particular, mains voltage sometimes affects the input impedance of plugged devices and its eventual generation of noise, what influences PLC channel response and received noise. The cyclic nature of PLC channels has been reported in the literature several times. One of the pioneer works in which this property is pointed out, though briefly is [16], others in which it has been covered with more detail are [17]–[23].

PLC channels exhibit also other kind of time variations due to the usage of appliances leading to relevant changes in the channel characteristics [24]. In [25], the relation between these long-term channel variations and the appearance of strong impulsive noise events is studied. Those are attributed to load changes in the power network, due to the connection and disconnection of appliances. However, such variations occur in a larger time scale (with respect to the signal modulation time scale) and not necessarily imply a clear periodicity (although tendencies along the day can be observed). Hence, their impact on communication systems design must be addressed with different strategies (e.g. re-configuring the link after detecting an important decrease on the received signal quality) that are out of the scope of this work.

The main contribution of this paper is to present general properties that cyclic signals and system exhibit but developed for the special case of PLC channels in which there is a dominant periodicity of the same period in both the CS random signals and LPTV systems. To the best authors knowledge, this approach has not been reported yet. Proofs for these properties are presented and examples of applications and practical considerations are given. A further, and reasonable, simplification of them that applies to PLC channels, derived from a slow variation LPTV model [21], completes the analysis. A review of current technology examples in PLC systems that have been designed taking into account some of these cyclic properties is included as well.

The paper is organized as follows. In the next section, the cyclic behavior of actual PLC channels is illustrated. The third section is devoted to review the characterization of cyclostationary random signals and to derive some mathematical relations useful for their analysis. The fourth section deals with LPTV systems and the study therein is focused in their filtering properties with CS signals. The next one presents a slow variation approximation of LPTV that is valid for PLC systems. Later, in section VI, a discussion is given on results from the literature where PLC channels cyclic behavior has been investigated. In section VII, a summary of common techniques employed in current PLC systems aware of this cyclic behavior are reviewed. Finally, in the last section, some conclusions are given.

II. ON THE CYCLIC NATURE OF PLC CHANNELS

In order to justify the usefulness of an analysis by means of cyclostationary random signals and LPTV systems, the physical principle underneath the cyclic behavior is firstly pointed out and, then, some results from the literature that validate this cyclic features are described.

A. RATIONALE FOR THE CYCLIC BEHAVIOR

Let us first focus on some physical properties of the medium. In particular, power networks involve many loads, like home appliances, whose characteristics have an important influence on the channel response. These loads commonly contain non-linear circuits in their power supplies (like diodes, thyristors, transient suppressors, etc.) that, when excited with a large signal of low frequency, the mains voltage, plus a small signal of high frequency, the communication signal, yield a time varying behavior from the communication system point of view [17]. As a result, the whole network that integrates the power cables, essentially with a linear time invariant behavior, and the time-varying loads, can be modeled as a LPTV system. A mathematical proof for this can be found in [21] (Appendix I.A).

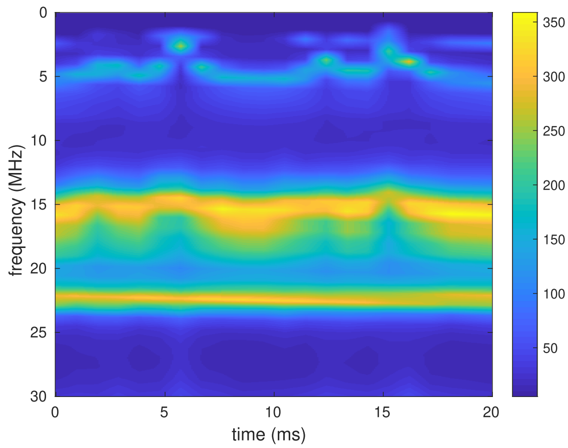

FIGURE 1. Measurement of the time varying impedance (real part in Ω) along a mains cycle of a microwave oven when it is working.

In this section, some examples of common appliances behavior are given for an illustrative purpose and to support the adequacy of a cyclic model. First, the input impedance taken for a microwave oven is shown in Fig.1, which has been estimated by means of measurements with a network analyzer synchronized with the mains period of 20 ms. As observed, the impedance exhibits a clear time variation related to the mains that is indeed frequency selective, i.e. there are different kind of variations at different bands. When such kind of impedance is connected to the power network, it acts as the load on a transmission line that determines to some extent the channel response. Depending on the selected frequency and where the transmitter and receiver are plugged, the influence on this response may be more or less intense.

Second, regarding the noise cyclostationarity, Fig.2 depicts the noise generated by a low-energy lamp. The power spectral density (PSD) estimated by means of signal processing of measurements synchronized with the mains period is shown. This plot serves as an example of commuted behavior in which two states appear twice along a mains period, i.e. the periodicity in this case is of 10 ms or with a 100 Hz fundamental frequency. When this kind of noise enters the power network, it is filtered by the system response and reaches

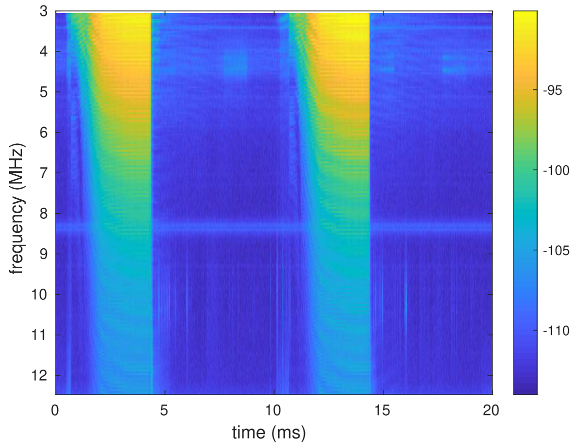

FIGURE 2. Measurement of the cyclostationary power spectral density (in dBm/kHz) along a mains cycle generated by a low-energy lamp.

the receiver, where the components both from the devices connected in the proximity and from the general distribution networks are superimposed.

Although we refer to T_0 as the mains period, often the fundamental period of the cyclic variation is half the mains period (10 ms or 8.33 ms depending on whether the mains frequency is 50 Hz or 60 Hz), due to a non-linear behavior of the loads (with a quadratic term or an absolute value).

B. EXPERIMENTAL EVIDENCES ON THE CYCLIC BEHAVIOR

The relevance of cyclic features in actual PLC channels depends on many factors. Being the appliances connected to the PLC network likely the most important one and, in particular, the built-in AC circuits (power supply, protection devices, electromagnetic interference filters, etc.). Appliances with low quality power supplies often influence significantly the PLC channel established over the network they are plugged in, enhancing the cyclic behavior. A significant work whose results validate these ideas is the one in [26] where, on one hand impedance measurements of some typical electrical loads that exhibit periodic variations are reported and, on the other hand, channel frequency responses of links at a small apartment containing such loads are presented as well. The study is carried out in a frequency band between 10 kHz and 30 MHz with a network analyzer. Variations larger than 5 dB along the mains cycle in the amplitude of the channel frequency response are observed at many frequencies (reaching 15 dB in certain cases). In addition, the apartment power network is simulated by means of transmission lines at which three time-varying impedance loads were connected. It is confirmed that the variation estimated by simulation in the channel frequency response is similar to the one obtained from measurements. Hence, it can be derived that the cables topology essentially acts as linear time invariant (LTI) system and the time variation is due to the loads behavior.

A measurement campaign over 50 links in power networks at homes, laboratories and small offices can be found in [17], [19], [21]. The set-up was designed to unveil the channel cyclic properties and was based on orthogonal frequency division multiplexing (OFDM) signal processing for broad band systems (from 1 to 30 MHz), synchronized with mains voltage. Illustrative results on the channel variation were presented with:

- spectrograms of the short-time Fourier transform of the channel response;
- excursion along the mains cycle, both of the amplitude and phase channel response at certain frequencies;
- estimated probability distributions of such frequency response excursion;
- estimated distributions of the frequency response rate of change in time, and of the channel Doppler spread, etc.

From those studies, the following conclusions can be summarized:

- the cyclic characteristics are mostly confined in the band below 10 MHz;
- the channel delay spread magnitude does not change too much along the mains cycle (which is beneficial to design the cyclic prefix of multi-carrier systems);
- changes of more than 6 dB of the channel frequency response amplitude over the mains cycle at many carriers were detected;
- in 40% of all the link carriers, variations along the mains cycle higher than 10% in amplitude were registered (which is important for channel estimation and equalization);
- the Doppler spread due to the time variation was estimated below 400 Hz for 90% of the measured channels being the median value 100 Hz.

Regarding the noise registered in the same trial, results were shown with:

- spectrograms of the short-term power spectral density (PSD) in the time-frequency plane;
- examples of average in time of these PSD (frequency selectivity);
- examples of average in frequency of these PSD (time selectivity);
- estimated distributions of the PSD excursion in time and of its rate of change in time, etc.

It can be highlighted that the cyclic noise:

- exhibits variations of more than 20 dB along the mains cycle at certain bands, and
- in the 30% of all link carriers, cyclic variations above 6 dB in the PSD were registered.

Other work in [27] described the implementation of a channel sounder for low voltage power lines, based on a FPGA board, valid for narrowband and broadband measurements. This tool allows registering channel responses and noise in a university building that exhibited relevant variations, especially in a frequency band between 10 kHz and 1 MHz. In the received noise, PSD excursions along mains cycle were higher than 20 dB, and higher than 10 dB in the channel response amplitude.

An interesting experiment was developed in [28], where some simple power networks were deployed in a laboratory, including a load with significant time variations: a cell phone charger. The analysis was in a frequency band up to 35 MHz and the authors showed that the amplitude of the channel response exhibited changes along the mains period between 10 and 25 dB at different frequencies. Commercial PLC modems, implementing OFDM technology, were tested on those networks and User Datagram Protocol (UDP) throughput differences larger than 30 Mb/s were obtained depending on whether the charger was connected or not. Also in [29], channel time variations are presented from measurements up to 50 MHz of realistic home power line networks. Examples with cyclic excursions of more than 10 dB in the amplitude are shown. In NB-PLC scenarios, experiments have been also carried out that present channel time variations [30].

In [31], noise measurements for narrow-band PLC (NB-PLC) systems (below 200 kHz) at different locations (office, laboratory, residential and industrial) are shown and cyclic features are analyzed. Variations in the PSD of more than 20 dB along the mains cycle are presented. The impulsive character of such variations and their magnitude was reported higher in the industrial location.

Once the presence of cyclic characteristics on PLC channels has been introduced, the next two sections are devoted to a formal treatment of cyclic signals and systems.

III. CYCLOSTATIONARY RANDOM SIGNALS

In the following, a notation in the continuous-time domain is employed because it is usually more intuitive, since mains voltage has an analog nature, although it can be straightforwardly translated to the discrete-time domain in which signal processing techniques are designed.

Just to fix notation, let us present the Fourier series (FS) equations for synthesis and analysis of periodical signals. $x(t)$ is a periodic signal in t with period T_0 when $x(t+T_0)=x(t)$ and, then, it can be represented as

$$x(t)=\sum_{\alpha=-\infty}^{+\infty} X^{\alpha} e^{j2\pi\alpha t/T_0}, \tag{1}$$

where the cyclic coefficients X^{α} of the FS,[1] can be computed by means of

$$X^{\alpha}=\frac{1}{T_0}\int_{-T_0/2}^{T_0/2} x(t)e^{-j2\pi\alpha t/T_0}\mathrm{d}t. \tag{2}$$

We are assuming that the series in (1) converges,[2] which is the case for many well-behaved real and complex functions $x(t)$ of interest.

A time-domain random signal $X(t)$ is referred to as strict-sense cyclostationary with period T_0, when its statistical properties are invariant after a time shift multiple of T_0, i.e. $X(t)$ and $X(t-nT_0)$ have identical statistics for all integer n [6]. A signal is cyclostationary (CS) in the wide-sense (what is assumed from now on) if both its mean and autocorrelation are periodical with any common period. In general terms, any non-stationary random signal with some periodical component, in its first or second moment, can be said to exhibit cyclostationarity, poly-cyclostationarity or almost-cyclostationarity [7], [10]. However, from now on, this paper deals with single-periodicity CS signals because they are the adequate to analyze PLC channels behavior.

A. MEAN, AUTOCORRELATION AND CYCLIC SPECTRUM

The mathematical expectation, expected value or *mean* of a CS random signal $X(t)$ with period T_0, is periodical in t, that is,

$$m_X(t)=E[X(t)]=E[X(t-nT_0)]=m_X(t-nT_0) \tag{3}$$

and, hence, it can be expanded with a FS

$$m_X(t)=\sum_{\alpha=-\infty}^{+\infty} m_X^{\alpha} e^{j2\pi\alpha t/T_0} \tag{4}$$

whose coefficients can be obtained from

$$m_X^{\alpha}=\frac{1}{T_0}\int_{-T_0/2}^{T_0/2} m_X(t)e^{-j2\pi\alpha t/T_0}\mathrm{d}t. \tag{5}$$

Analogously, the *autocorrelation*[3] of CS signal $X(t)$ is also periodical in t with period T_0

$$\begin{aligned} R_X(t,t+u) &= E\left[X^*(t)X(t+u)\right] \\ &= E\left[X^*(t-nT_0)X(t-nT_0+u)\right] \\ &= R_X(t-nT_0,t-nT_0+u), \end{aligned} \tag{6}$$

and can be expanded with a FS

$$R_X(t,t+u)=\sum_{\alpha=-\infty}^{+\infty} R_X^{\alpha}(u)e^{j2\pi\alpha t/T_0}; \tag{7}$$

from the corresponding coefficients given by

$$R_X^{\alpha}(u)=\frac{1}{T_0}\int_{-T_0/2}^{T_0/2} R_X(t,t+u)e^{-j2\pi\alpha t/T_0}\mathrm{d}t. \tag{8}$$

It is worth to observe some properties of FS coefficients. In particular, zero-order coefficient is the time average of autocorrelation function along a mains cycle

$$R_X^{0}(u)=\frac{1}{T_0}\int_{-T_0/2}^{T_0/2} R_X(t,t+u)\mathrm{d}t. \tag{9}$$

It is straightforward to test the time independence of this average. Let us define a new random signal $Y(t)=X(t+t')$,

[1] Note that super-indices in greek letters, when applied to a function or variable, must be interpreted as the order of the series coefficient and not as exponentiation.

[2] That is, that Dirichlet conditions are satisfied.

[3] In other works devoted to CS signals, like the ones by Gardner [6], a definition of the autocorrelation function with a symmetric time shift is given. This leads to different formulation of relations in the frequency domain like (22).

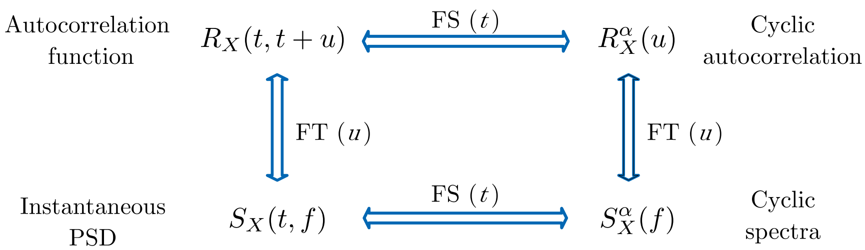

FIGURE 3. Relations by means of FS and FT pairs of the autocorrelation and power spectral density functions for CS signals.

which will be clearly also a CS signal, and the cyclic coefficients of the autocorrelation function are

$$R_Y^\alpha(u) = R_X^\alpha(u)e^{j2\pi\alpha t'/T_0}. \tag{10}$$

whose particular case for $\alpha = 0$ proves the time independence. Otherwise, higher-order coefficients do not hold such time-shifting invariance but exhibit a cyclic variation, sinusoidal of frequency α/T_0. A fact that yields $R_X^\alpha(u)$ to be referred to as the *cyclic autocorrelation* and the α-th term quantifies the contribution at each harmonic of the fundamental frequency. A classical method to test whether a random process is CS for a certain period or not, consists in evaluating if its cyclic autocorrelation by means of (8) presents any non null term for $\alpha \neq 0$.

In a similar manner to the Wiener-Khinchin theorem for wide-sense stationary signals, the Fourier transform (FT) can be applied to the autocorrelation function of CS signals to define the instantaneous power spectral density (IPSD) as

$$S_X(t,f) = \int_{-\infty}^{+\infty} R_X(t,t+u)e^{-j2\pi fu}\mathrm{d}u. \tag{11}$$

The IPSD inherits the periodicity in t of the autocorrelation function and, hence can be expanded by means of the synthesis equation of the FS (1),

$$S_X(t,f) = \sum_{\alpha=-\infty}^{+\infty} S_X^\alpha(f)e^{j2\pi\alpha t/T_0}, \tag{12}$$

being the expression to calculate these FS coefficients (2)

$$S_X^\alpha(f) = \frac{1}{T_0}\int_{-T_0/2}^{T_0/2} S_X(t,f)e^{-j2\pi\alpha t/T_0}\mathrm{d}t, \tag{13}$$

also referred to as the *cyclic spectra* of the CS signal which can be related to the cyclic autocorrelation by substituting (7) in (11) and taking this result to (12)

$$S_X^\alpha(f) = \int_{-\infty}^{+\infty} R_X^\alpha(u)e^{-j2\pi fu}\mathrm{d}u. \tag{14}$$

The case for $\alpha = 0$, makes (13) to represent the time average of the IPSD along a mains cycle. As seen later, $S_X^0(f)$ usually denotes the PSD of the CS signal. This way, the basic set of expressions between the different autocorrelation functions and power spectral densities related by FS and FT pairs have been presented, what is illustrated in Fig.3.

B. POWER SPECTRAL DENSITY OF CYCLOSTATIONARY SIGNALS

The PSD is naturally associated to the measure of the signal power in a certain frequency band, which can be estimated by means of a spectrum analyzer. That is an intuitive idea valid for the study of stationary signals but needs some adjustment for CS signals. In the latter case, a random time shift can be introduced to the CS signal to break its periodicity and reach such power measure.[4] A CS signal $X(t)$ yields a stationary wide-sense process $Y(t) = X(t-\tau)$, when τ is a random variable uniformly distributed in the interval $[0, T_0]$ independent of $X(t)$. It is straightforward to prove that the mean of $Y(t)$ is the time average of the mean of $X(t)$ and independent of t

$$\begin{aligned} m_Y &= E_{X(t)}\left[E_\tau[X(t-\tau)|\tau]\right] \\ &= E\left[\int_{-\infty}^{\infty} X(t-\tau)f_\tau(\tau)\mathrm{d}\tau\right] \\ &= E\left[\int_0^{T_0} X(t-\tau)\frac{1}{T_0}\mathrm{d}\tau\right] = \frac{1}{T_0}\int_0^{T_0} E[X(t-\tau)]\mathrm{d}\tau \\ &= \frac{1}{T_0}\int_0^{T_0} m_X(t-\tau)\mathrm{d}\tau = m_X^0 \end{aligned} \tag{15}$$

Analogously, the autocorrelation function of $Y(t)$, is the time average of $R_X(t,t+u)$

$$R_Y(u) = E[R_X(t-\tau, t-\tau+u)] = R_X^0(u). \tag{16}$$

Using the transform pair in (14), the relation $S_Y(f) = S_X^0(f)$ can be derived, and the mean power definition of a random signal from its PSD [32] leads to

$$E[|Y^2(t)|] = \int_{-\infty}^{+\infty} S_X^0(f)\mathrm{d}f, \tag{17}$$

[4]Phase randomization of a CS, or time-averaging its probability distribution, can result in hidden cyclostationarity, changing it to a stationary random signal [7]. For instance, this procedure is commonly used to derive the PSD of a pulse-amplitude modulated digital signal, which is CS in essence (provided that the symbol sequence is stationary) but is converted into stationary with a random time shift independent of its symbol period [6].

which represents the mean power of $Y(t)$. Hence, $S_X^0(f)$, the zero-order cyclic spectrum of $X(t)$, constitutes the PSD of $Y(t)$. In practice, this would be the measure obtained with a spectrum analyzer when excited with the CS signal $X(t)$ (provided that the sweep time is not synchronized with the signal period and several traces or registers are averaged). For that reason, $S_X^0(f)$ is usually referred to as the PSD of the CS signal.

C. CYCLOSTATIONARY SIGNALS AND LTI SYSTEMS

In this section, the filtering of CS signals through LTI systems is shortly described. Intuition says that a time-invariant operation applied to a CS process with period T_0 will yield a new CS process as shown in Fig.4.

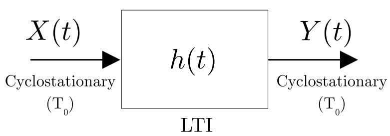

FIGURE 4. **LTI filtering of a CS signal.**

Let us recall the input-output relation of a LTI filter, determined by the convolution integral of the excitation signal with its impulse response $h(t)$:

$$Y(t) = h(t) * X(t) = \int_{-\infty}^{+\infty} h(t-u)X(u)\mathrm{d}u. \tag{18}$$

Property 1: When exciting a LTI system of frequency response $H(f)$ with a CS input signal $X(t)$, the cyclic coefficients of the output signal mean are given by

$$m_Y^\alpha = m_X^\alpha \cdot H\left(\frac{\alpha}{T_0}\right). \tag{19}$$

Proof: The mean of the output signal is

$$m_Y(t) = \int_{-\infty}^{+\infty} h(t-u)m_X(u)\mathrm{d}u = \int_{-\infty}^{+\infty} h(s)m_X(t-s)ds. \tag{20}$$

This expression is periodical in time with period T_0 like $m_X(t)$ is and, hence, it can be expanded in a FS with coefficients given by (5)

$$m_Y^\alpha = \frac{1}{T_0}\int_{-T_0/2}^{T_0/2}\int_{-\infty}^{+\infty} h(s)m_X(t-s)ds\, e^{-j2\pi\alpha t/T_0}\mathrm{d}t. \tag{21}$$

After simple algebraic manipulations, it can be derived that

$$\begin{aligned} m_Y^\alpha &= \frac{1}{T_0}\int_{-T_0/2}^{T_0/2} m_X(t-s)\, e^{-j2\pi\alpha(t-s)/T_0}\,\mathrm{d}t \\ &\quad \cdot \int_{-\infty}^{+\infty} h(s)e^{-j2\pi\alpha s/T_0}\mathrm{d}s \\ &= m_X^\alpha \int_{-\infty}^{+\infty} h(s)e^{-j2\pi\alpha s/T_0}ds = m_X^\alpha \cdot H\left(\frac{\alpha}{T_0}\right), \end{aligned}$$

which completes the proof. ■

Property 2: When exciting a LTI system of frequency response $H(f)$ with a CS input signal $X(t)$, the cyclic spectra of the output signal can be obtained as[5]

$$S_Y^\alpha(f) = H^*\left(f - \frac{\alpha}{T_0}\right)H(f)S_X^\alpha(f). \tag{22}$$

Proof: From the input signal autocorrelation function definition in (6) and, using (17), we have

$$\begin{aligned} R_Y(t, t+u) &= E[Y^*(t)Y(t+u)] \\ &= E[\int_{-\infty}^{+\infty} h^*(t-s)X^*(s)ds \int_{-\infty}^{+\infty} h(t+u-v)X(v)\mathrm{d}v] \\ &= \iint_{-\infty}^{+\infty} h^*(t-s)h(t+u-v)E[X^*(s)X(v)]\mathrm{d}s\mathrm{d}v \\ &= \iint_{-\infty}^{+\infty} h^*(t-s)h(t+u-v)R_X(s,v)\mathrm{d}s\mathrm{d}v. \end{aligned} \tag{23}$$

By changing the integration variable, $v = s + w$, the so-far used expression of the input autocorrelation is obtained

$$\begin{aligned} R_Y(t, t+u) = \iint_{-\infty}^{+\infty} & h^*(t-s)h(t+u-s-w) \\ & \cdot R_X(s, s+w)\mathrm{d}s\mathrm{d}w, \end{aligned} \tag{24}$$

which lead to a valuable intermediate result that connects the autocorrelation of the output and input signals. Taking into account the CS nature of $X(t)$, function $R_X(s, s+w)$ can be expanded in a FS by means of (7) to yield (25), where a change in the integration variable $r = t - s$ has been made.

$$\begin{aligned} R_Y(t, t+u) &= \iint_{-\infty}^{+\infty} h^*(t-s)h(t+u-s-w) \\ &\quad \times \sum_{\alpha=-\infty}^{+\infty} R_X^\alpha(w)e^{j2\pi\alpha s/T_0}\mathrm{d}s\mathrm{d}w \\ &= \iint_{-\infty}^{+\infty} h^*(r)h(r+u-w) \\ &\quad \times \sum_{\alpha=-\infty}^{+\infty} R_X^\alpha(w)e^{j2\pi\alpha t/T_0}e^{-j2\pi\alpha r/T_0}\mathrm{d}r\mathrm{d}w \\ &= \sum_{\alpha=-\infty}^{+\infty}\iint_{-\infty}^{+\infty} h^*(r)h(r+u-w)e^{-j2\pi\alpha r/T_0} \\ &\quad \times R_X^\alpha(w)\mathrm{d}r\mathrm{d}w \cdot e^{j2\pi\alpha t/T_0}. \end{aligned} \tag{25}$$

[5] As mentioned before, the definition of the autocorrelation function with an asymmetrical time shift in (6), leads to an asymmetrical shift in the frequency response terms in (22). But, this way, harmonic shifts with the fundamental frequency are more clear than in other analogous expressions in the literature [6].

The expression in (25) has the form of a FS of the output signal autocorrelation, $R_Y(s, s+w)$, whose coefficients are

$$R_Y^{\alpha}(u)=\iint\limits_{-\infty}^{+\infty} h_{\alpha}(r)h(r+u-w)e^{-j2\pi\alpha r/T_0}R_X^{\alpha}(w)\mathrm{d}r\mathrm{d}w. \quad (26)$$

For the sake of clarity, in this equation an auxiliary function $h_{\alpha}(t)$, with the form of an impulse response, has been defined as

$$h_{\alpha}(t) = h^*(t)e^{-j2\pi\alpha t/T_0}. \quad (27)$$

Applying signal convolution operation, successively to the two integration variables, (26) can be modified to

$$\begin{aligned} R_Y^{\alpha}(u) &= \int\limits_{-\infty}^{+\infty} [h_{\alpha}(w-u) * h(u-w)]R_X^{\alpha}(u)\mathrm{d}u \\ &= h_{\alpha}(-u) * h(u) * R_X^{\alpha}(w), \end{aligned} \quad (28)$$

which relates the cyclic coefficients of the output and input signals autocorrelation. Since PSD is the FT of an stationary process autocorrelation, (14) can be used to express the output cyclic spectrum in terms of the input ones

$$S_Y^{\alpha}(f) = H_{\alpha}(-f)H(f)S_X^{\alpha}(f). \quad (29)$$

Finally, by substituting in (29) the FT of $h_{\alpha}(t)$ in (27)

$$\begin{aligned} H_{\alpha}(f) &= H^*(-f) * \delta\left(f+\frac{\alpha}{T_0}\right) \\ H_{\alpha}(-f) &= H^*(f) * \delta\left(f-\frac{\alpha}{T_0}\right) = H^*\left(f-\frac{\alpha}{T_0}\right), \end{aligned} \quad (30)$$

the proof of (22) is completed. ■

A Practical example: PLC channel response characterization by means of a spectrum analyzer. Previous result helps to describe some effects of applying a CS signal, like noise of PLC channels, to a spectrum analyzer. Basically, this instrument works in a similar manner to a superheterodyne receiver, i.e., the convolution of excitation signal with the impulse response of a tuned filter (let us consider it almost ideal, with an almost rectangular frequency response) whose central frequency, f_c, is shifted during the sweep time. Such impulse response would be

$$h(t)\cdot e^{j2\pi f_c t} \overset{FT}{\Leftrightarrow} H(f-f_c). \quad (31)$$

Despite the fact that a spectrum analyzer is designed to operate with stationary signals, this instrument can serve to estimate the instantaneous PSD of the power line CS noise. To do so, the sweep time must be synchronized with the mains period, that is, each frequency step must go after one period T_0. Hence the entire sweep time would be as long as T_0 times the desired number of frequency samples. By triggering the sweep at a certain initial mains cycle time phase $t = t_i$ and setting the resolution bandwidth higher that the fundamental mains frequency $1/T_0$, it can be assumed that $H(f-\frac{\alpha}{T_0}) \simeq H(f)$ and the following approximation holds

$$\begin{aligned} S_Y(t_i, f) &= \sum_{\alpha=-\infty}^{+\infty} H^*(f-\frac{\alpha}{T_0})H(f)S_X^{\alpha}(f)e^{j2\pi\alpha t_i/T_0} \\ &\simeq |H(f)|^2 \sum_{\alpha=-\infty}^{+\infty} S_X^{\alpha}(f)e^{j2\pi\alpha t_i/T_0} \\ &\simeq |H(f)|^2 \cdot S_X(t_i, f). \end{aligned} \quad (32)$$

This expression constitutes an estimation of the noise IPSD of $S_Y(t, f)$ valued for $t = t_i$. By repeating the procedure for different t_i to walk the whole period T_0, the time variation of $S_Y(t, f)$ can be sampled reasonably.

This procedure may be used as an experimental method to verify whether a signal (which we know is stationary or at least CS) exhibits cyclostationarity for a certain period T_0 or not: it is enough to observe two analyzer measurement (with the sweep synchronized with T_0) triggered at different time phases $t_i \neq t_k$. If the two measurements are not the same, the signal is CS with period T_0. Obviously, there are other procedures more reliable and sophisticated to identify CS signals by means of data acquisition devices and signal processing algorithms.

Let us review some interesting consequences derived from the previous properties. When a CS process is filtered by an low-pass LTI system (like depicted in Fig.4), frequency selective enough so that $H(f) = 0\ \forall f > 1/T_0$, then

$$m_Y^{\alpha} = 0\ \forall\ \alpha \neq 0\ \Rightarrow\ m_Y = H(0)m_X, \quad (33)$$

the output process mean is invariant with t and, also,

$$S_Y^{\alpha}(f) = 0\ \forall\ \alpha \neq 0\ \Rightarrow\ S_Y(f) = |H(f)|^2 S_X^0(f), \quad (34)$$

its IPSD does not exhibit cyclic variation either (i.e. it is no longer instantaneous), hence, it is stationary. The LTI system is performing a severe averaging to the excitation signal that exceeds the CS period.

D. DECOMPOSITION OF A CS SIGNAL AS THE COMBINATION OF A SET OF STATIONARY SIGNALS

In the analysis of CS signals it is often helpful to decompose them in a set of stationary signals. There are different strategies to do so, among which the synchronized sampling with the CS signal period and the sub-band filtering can be mentioned.

1) DECOMPOSITION IN TIME BY SYNCHRONIZED SAMPLING

This is likely the most straightforward technique [12]. Let us assume that a CS signal $X(t)$ is sampled in time by taking L samples uniformly spaced each T_s over an interval equals to T_0, the cyclic variation period, i.e. $L = T_0/T_s$. Let n be the discrete-time index $t = nT_s$, which may be expressed as $n = mL + \ell$, being[6] $m = \lfloor t/T_0 \rfloor = \lfloor (nT_s)/(LT_s) \rfloor = \lfloor n/L \rfloor$

[6] $\lfloor x \rfloor$ denotes the integer part of x.

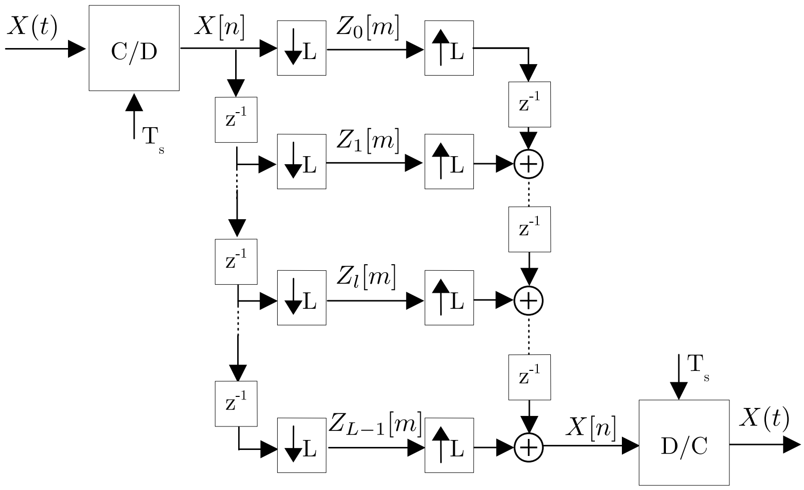

FIGURE 5. Decomposition of a CS signal in a set of stationary discrete-time signals and the subsequent synthesis.

and $\ell = 0, 1 ... L-1$ is the sampling-time phase in a cycle (otherwise said: $\ell = \text{mod}(n, L)$).

Consequently, L discrete-time signal are obtained, one for each sampling-time phase, which represent a decimation of $X(t)$ and will be denoted

$$Z_\ell[m] = X(mT_0 + \ell T_s) = X(\lfloor t/T_0 \rfloor T_0 + \ell T_s). \quad (35)$$

By combining such signals, a discrete-time version of the original process can be synthesized as

$$X[n]=X(nT_s)=\sum_{\ell=0}^{L-1} Z_\ell[\lfloor n/L \rfloor] \cdot \delta[n - \lfloor n/L \rfloor L - \ell]. \quad (36)$$

The relation between discrete-time and continuous-time signals, provided that the sampling rate is higher enough to avoid aliasing, is [33]

$$X(t) = \sum_{n=-\infty}^{+\infty} X[n] \cdot \text{sinc}(t/T_s - n). \quad (37)$$

By means of (36) and (37) the initial signal $X(t)$ can be expressed in terms of the discrete-time signals $Z_\ell[m]$. This analysis procedure is illustrated in Fig.5, where interpolation of the L decimation phases to synthesize the original random process is shown as well.

To verify the stationary character of $Z_\ell[m]$, let us consider them as a sampled version of $X(t)$ with a sampling period equals to the CS cycle, $T_s' = T_0$, but with different time initial phases $t_\ell = \ell T_s$,

$$Z_\ell[m] = X(mT_0 + t_\ell). \quad (38)$$

By reminding that $X(t)$ is a CS signal and by using (3),

$$\begin{aligned} m_{Z_\ell} &= E[X(mT_0 + t_\ell)] = E[X(t_\ell)] = m_X(t_\ell) \\ &= \sum_{\alpha=-\infty}^{+\infty} m_X^\alpha e^{j2\pi\alpha t_\ell/T_0}; \end{aligned} \quad (39)$$

so the mean of $Z_\ell[m]$ only depends on t_ℓ and not on m. Its autocorrelation is also a function only of t_l, as can be seen by using (6),

$$R_{Z_\ell}[p] = R_x(t_\ell, t_\ell + pT_0) = \sum_{\alpha=-\infty}^{+\infty} R_X^\alpha(pT_0) e^{j2\pi\alpha t_\ell/T_0}. \quad (40)$$

Since, neither m_{Z_ℓ} nor $R_{Z_\ell}[p]$ depends on m, and neither on t, they are stationary random signals. In fact, many times, $R_{Z_\ell}[p]$ is zero for $p > 0$ if $R_x(\tau)$ is not very long (at least compared with T_0), which is the common case in power line communications.

2) DECOMPOSITION IN FREQUENCY BY SUB-BAND FILTERING

An alternative procedure to decompose a CS signal in stationary ones in the sub-band filtering [6], which is based on the CS signal $X(t)$, with cycle period T_0, representation with a set of stationary signals $A^\alpha(t)$, $\alpha = 0, \pm 1, \pm 2, ...$, as

$$X(t) = \sum_{\alpha=-\infty}^{+\infty} A^\alpha(t) e^{j2\pi\alpha t/T_0}, \quad (41)$$

being

$$A^\alpha(t) = [X(t)e^{-j2\pi\alpha t/T_0}] * w(t), \quad (42)$$

where the ancillary function

$$w(t) = \frac{1}{T_0}\text{sinc}(t/T_0) \quad (43)$$

has been included. The convolution operation (42) represents a selective filtering that, by reminding (33) and (34), supports the stationarity of the output signal. Equation (42) can be reformulated to get closer to a filter bank implementation as depicted in Fig.6,

$$A^\alpha(t) = X(t) * [w(t)e^{j2\pi\alpha t/T_0}]e^{-j2\pi\alpha t/T_0}. \quad (44)$$

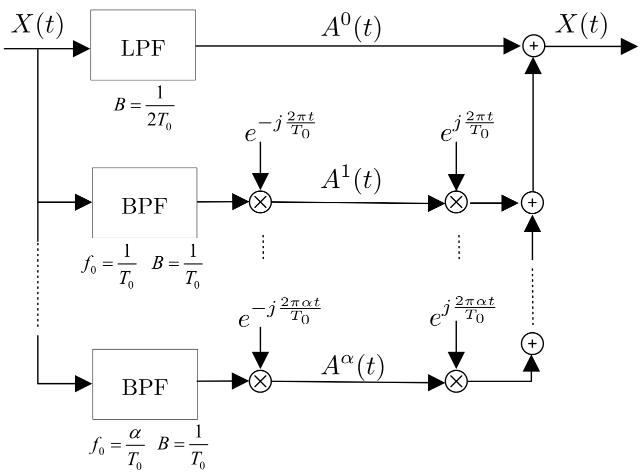

FIGURE 6. Decomposition of a CS signal in a set of stationary signals by means of a filter bank and its subsequent synthesis.

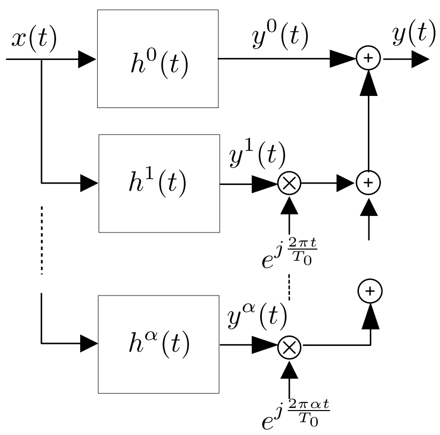

FIGURE 7. Decomposition of a LPTV system in a set of synchronized LTI systems.

This expression can be translated to the frequency domain and it represents a filtering in bands of $1/T_0$ width, centered at harmonics of the fundamental frequency $1/T_0$, employing ideal filters with rectangular frequency response like

$$H^\alpha(f) = W\left(f - \frac{\alpha}{T_0}\right) = \begin{cases} 1, & si\ |f - \frac{\alpha}{T_0}| < \frac{1}{2T_0} \\ 0, & \text{otherwise} \end{cases} \qquad (45)$$

If the input random process is band-limited, it will exhibit a finite number of such stationary components that are referred to as the harmonic-series representation and can be interpreted as a frequency division demultiplexing [1], [6].

This approach has been followed in PLC as a way to synthesize realizations of CS random signals to be employed in simulations. For instance, in the IEEE Standard for ''Low Frequency (less than 500 kHz) Narrow Band Power Line Communications for Smart Grid Applications'' [34] a procedure to generate CS noise similar to this one is described.

IV. LINEAR PERIODICALLY TIME-VARYING SYSTEMS

A linear Time-Variant (LTV) can be described by means of its input-output relation

$$y(t) = \int_{-\infty}^{+\infty} h(t, u)x(u)\mathrm{d}u \qquad (46)$$

where $h(t, u)$ represent the transformation kernel and is the *system impulse response*, measured at instant t when excitation is applied at instant u [6], [12]. Simplification to $h(t, u) = h(t - u, 0) = h(t - u)$ corresponds to the particular case of Linear Time-Invariant (LTI) systems. If the LTV impulse response is periodical in time with period T_0, that is,

$$h(t - nT_0, t - nT_0 - \tau) = h(t, t - \tau), \qquad (47)$$

the system is referred to as Linear Periodically Time-Varying (LPTV).[7]

[7] Notation $h(t, t-\tau)$ highlights the delay between excitation and response.

A. TIME AND FREQUENCY RESPONSE OF LPTV SYSTEMS

LPTV system response can be expanded with a FS in t, by using the synthesis equation

$$h(t, t - \tau) = \sum_{\alpha=-\infty}^{+\infty} h^\alpha(\tau)e^{j2\pi\alpha t/T_0}, \qquad (48)$$

and the *cyclic impulse response* coefficients can be calculated with the analysis equation of the FS,

$$h^\alpha(\tau) = \frac{1}{T_0}\int_{-T_0/2}^{T_0/2} h(t, t - \tau)e^{-j2\pi\alpha t/T_0}\mathrm{d}t. \qquad (49)$$

From these relations, a decomposition of the LPTV system in a set of LTI ones can be derived in a natural manner, by introducing (48) in the system impulse response (46)

$$y(t) = \sum_{\alpha=-\infty}^{+\infty}\int_{-\infty}^{\infty} h^\alpha(\tau)x(t - \tau)\mathrm{d}\tau \cdot e^{j2\pi\alpha t/T_0}, \qquad (50)$$

which yields, at the output, a superposition of LTI filter responses $h^\alpha(t)$ that have been modulated with harmonics of the fundamental cyclic frequency $1/T_0$

$$y(t) = \sum_{\alpha=-\infty}^{+\infty} y^\alpha(t)e^{j2\pi\alpha t/T_0}. \qquad (51)$$

This latter equation can be transformed to the frequency domain to reach

$$\begin{aligned} Y(f) &= \int_{-\infty}^{+\infty}\sum_{\alpha=-\infty}^{+\infty} y^\alpha(t)e^{j2\pi\alpha t/T_0}e^{-j2\pi ft}\mathrm{d}t \\ &= \sum_{\alpha=-\infty}^{+\infty}\int_{-\infty}^{+\infty} y^\alpha(t)e^{-j2\pi(f-\frac{\alpha}{T_0})t}\mathrm{d}t \\ &= \sum_{\alpha=-\infty}^{+\infty} Y^\alpha\left(f - \frac{\alpha}{T_0}\right), \end{aligned} \qquad (52)$$

which can be interpreted as the superposition of LTI filtered signals FT, but frequency shifted. This represents a spectral

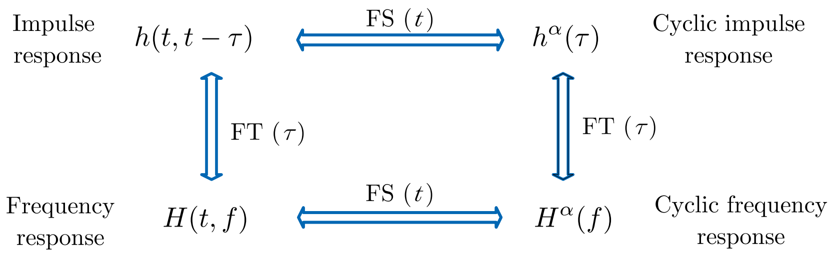

FIGURE 8. Relations by means of transform pairs between LPTV system responses and its cyclic coefficients.

broadening of the input signal (a feature of all LTV systems), whose magnitude depends on the cyclic period, the input signal bandwidth and the LTI filters' one.

In OFDM signals, this broadening represents an intercarrier interference (ICI) term that degrades the system performance, which cannot be compensated for by increasing the number of subcarriers as in LTI channels. This limit of the maximum number of subcarriers of the OFDM system (or equivalently, the symbol length) has been evaluated for PLC systems in [35].

In Fig.7, this equivalence of the LPTV system is described, with a structure that corresponds to the so-called frequency-shift (FRESH) filtering implementation [36].

The frequency response of a LPTV is defined as the FT in the τ variable of the impulse response, that is,

$$H(t,f) = \int_{-\infty}^{+\infty} h(t,t-\tau)e^{-j2\pi f\tau}\mathrm{d}\tau, \tag{53}$$

which is a periodic function in t with period T_0, since $h(t,t-\tau)$ is, i.e.

$$H(t-nT_0,f) = H(t,f). \tag{54}$$

Hence, the frequency response can be also expanded in FS, by substituting (48) in (53) to obtain

$$H(t,f) = \sum_{-\infty}^{+\infty} H^{\alpha}(f)e^{j2\pi\alpha t/T_0}, \tag{55}$$

whose coefficients, referred to as the *cyclic frequency response* are given by

$$H^{\alpha}(f) = \frac{1}{T_0}\int_{-T_0/2}^{T_0/2} H(t,f)e^{-j2\pi\alpha t/T_0}\mathrm{d}t. \tag{56}$$

Property 3: The output signal of a LPTV system when excited with an input signal $x(t)$ can be expressed in terms of the system cyclic response as

$$Y(f) = \sum_{\alpha=-\infty}^{+\infty} H^{\alpha}\left(f-\frac{\alpha}{T_0}\right)X\left(f-\frac{\alpha}{T_0}\right). \tag{57}$$

Proof: By virtue of (50) and (51), the identity $y^{\alpha}(t) = h^{\alpha}(t)*x(t)$ appears that, transformed to the frequency domain leads to the relation between the input and output cyclic coefficients

$$Y^{\alpha}(f) = H^{\alpha}(f)\cdot X(f). \tag{58}$$

With this latter expression and recalling (52), the spectral broadening that LPTV filtering causes on the input signal in terms of the cyclic frequency response can be derived as expressed in (57), what concludes the proof. ■

Finally, a diagram with the different relations of SF and FT pairs that can be established between the cyclic responses, in the time and frequency domains and their respective coefficients, is shown in Fig.8.

B. SYNTHESIS OF CS BY FILTERING STATIONARY SIGNALS THROUGH LPTV SYSTEMS

It is of particular interest to explore some relations between CS signals and LPTV transformations. In fact, as it will be demonstrated below, a procedure to obtain CS is by LPTV filtering of stationary random signals (as depicted in Fig.9); which constitutes an alternative method of CS signals synthesis that is complementary to the mentioned ones in the previous section.

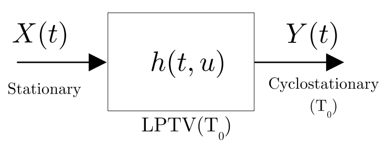

FIGURE 9. LPTV filtering of a stationary random signal.

Let us consider a stationary random process $X(t)$ that excites a LPTV system whose response can be reformulated from (46), for realizations of the signal input and output,

$$y(t) = \int_{-\infty}^{+\infty} h(t,t-u)x(t-u)\mathrm{d}u. \tag{59}$$

$$S_Y^{\alpha}(f) = \sum_{\beta=-\infty}^{+\infty}\left[H^{\beta}\left(f-\frac{\alpha+\beta}{T_0}\right)\right]^{*} H^{\alpha+\beta} \times \left(f-\frac{\alpha+\beta}{T_0}\right)S_X\left(f-\frac{\alpha+\beta}{T_0}\right) \tag{60}$$

Property 4: When exciting a LPTV system with a stationary signal $X(t)$, a CS signal is obtained at the output, whose mean is periodic with cyclic coefficients given by

$$m_Y^\alpha = m_X H^\alpha(0), \tag{61}$$

where $H^\alpha(f)$ are the system cyclic frequency response.

Proof: Let us begin with the calculation of the output signal mean by applying latter equation to the definition,

$$\begin{aligned} E[Y(t)] &= E[\int_{-\infty}^{+\infty} h(t, t-u)X(t-u)\mathrm{d}u] \\ &= \int_{-\infty}^{+\infty} h(t, t-u)E[X(t-u)]\mathrm{d}u. \end{aligned} \tag{62}$$

Stationary condition of $X(t)$ yields its mean to be independent of time, so $E[X(t)] = m_X$, and (62) turns into

$$\begin{aligned} m_Y(t) &= m_X \int_{-\infty}^{+\infty} h(t, t-u)\mathrm{d}u = m_X H(t, 0) \\ &= m_X \sum_{\alpha=-\infty}^{+\infty} H^\alpha(0)e^{j2\pi\alpha t/T_0}; \end{aligned} \tag{63}$$

where definition of the system frequency response in (53) and its SF expansion (55) have been used. By identifying this result with the cyclic decomposition of the mean in (4), (61) can be reached, concluding the proof. ■

Property 5: When a stationary signal excites a LPTV system, the cyclic spectra of the CS signal at the output can be expressed in terms of the input signal PSD and the system cyclic response as in (60).

Proof: To obtain the output signal autocorrelation, defined in (6) and reformulated here for clarity as

$$R_Y(t, t+u) = E[Y^*(t)Y(t+u)], \tag{64}$$

with the help of (59) and considering the periodic response of the system, the following algebraic manipulations are possible

$$\begin{aligned} Y^*(t) &= \int_{-\infty}^{+\infty} h^*(t, t-s)X^*(t-s)\mathrm{d}s \\ &= \int_{-\infty}^{+\infty} \sum_{\beta=-\infty}^{+\infty} [h^\beta(s)]^* e^{-j2\pi\beta t/T_0} X^*(t-s)\mathrm{d}s; \end{aligned} \tag{65}$$

and

$$\begin{aligned} Y(t+u) &= \int_{-\infty}^{+\infty} h(t+u, t+u-v)X(t+u-v)\mathrm{d}v \\ &= \int_{-\infty}^{+\infty} \sum_{\gamma=-\infty}^{+\infty} h^\gamma(v)e^{j2\pi\gamma(t+u)/T_0}X(t+u-v)\mathrm{d}v. \end{aligned} \tag{66}$$

Leading to

$$\begin{aligned} R_Y(t, t+u) &= E[\sum_{\beta=-\infty}^{+\infty} \sum_{\gamma=-\infty}^{+\infty} e^{-j2\pi\beta t/T_0} e^{j2\pi\gamma(t+u)/T_0} \\ &\quad \cdot \iint_{-\infty}^{+\infty} [h^\beta(s)]^* h^\gamma(v)X^*(t-s)X(t+u-v)\mathrm{d}s\mathrm{d}v] \\ &= \sum_{\beta=-\infty}^{+\infty} \sum_{\gamma=-\infty}^{+\infty} e^{j2\pi(\gamma-\beta)t/T_0} e^{j2\pi\gamma u/T_0} \\ &\quad \cdot \iint_{-\infty}^{+\infty} [h^\beta(s)]^* h^\gamma(v)R_X(u+s-v)\mathrm{d}s\mathrm{d}v, \end{aligned} \tag{67}$$

where input stationarity has been used to denote

$$R_X(t, t+u) = R_X(u). \tag{68}$$

By recalling convolution integral properties, (67) can be simplified to

$$\begin{aligned} R_Y(t, t+u) = \sum_{\beta=-\infty}^{+\infty} \sum_{\gamma=-\infty}^{+\infty} & e^{j2\pi(\gamma-\beta)t/T_0} e^{j2\pi\gamma u/T_0} \\ & \cdot \{[h^\beta(-u)]^* * h^\gamma(u) * R_X(u)\}. \end{aligned} \tag{69}$$

After reordering terms and making a change of sum indexes $\alpha = \gamma - \beta$, it yields to

$$\begin{aligned} R_Y(t, t+u) = \sum_{\alpha=-\infty}^{+\infty} \sum_{\beta=-\infty}^{+\infty} & e^{j2\pi(\alpha+\beta)u/T_0} \\ & \cdot \{[h^\beta(-u)]^* * h^{\alpha+\beta}(u) * R_X(u)\} e^{j2\pi\alpha t/T_0}, \end{aligned} \tag{70}$$

and by identifying terms in the FS expansion of the cyclic autocorrelation, see (7), it leads to

$$\begin{aligned} R_Y^\alpha(u) = \sum_{\beta=-\infty}^{+\infty} & e^{j2\pi(\alpha+\beta)u/T_0} \\ & \cdot \{[h^\beta(-u)]^* * h^{\alpha+\beta}(u) * R_X(u)\}, \end{aligned} \tag{71}$$

which defines the relation between the cyclic coefficients of the CS output signal autocorrelation with the input signal autocorrelation and the LPTV system response. Since cyclic spectra $S_Y^\alpha(f)$ form a transform pair with the autocorrelation cyclic coefficients $R_Y^\alpha(u)$, it is possible to obtain them by applying some FT properties and simple algebra to derive (60). ■

In the particular case of $\alpha = 0$, the PSD of the CS output process $Y(t)$ can be calculated (recall (16)):

$$S_Y^0(f) = \sum_{\beta=-\infty}^{+\infty} \left| H^\beta\left(f - \frac{\beta}{T_0}\right)\right|^2 S_X\left(f - \frac{\beta}{T_0}\right). \tag{72}$$

This expression represents the counterpart to LTI filtering of random stationary signals $S_Y(f) = |H(f)|^2 S_X(f)$. Finally, it is worth mentioning that the presented relations describe a method to create CS signals from stationary ones.

A practical example: power line modulation. LPTV filtering of CS signals is a common phenomenon in communication systems, for instance, it appears in modulation sub-systems. In PLC, it was pointed out by measurements of the pioneer work by Chan and Donaldson [16], in which authors try to characterize the attenuation of power lines and a variation versus time appeared as a periodical fading, with 5 dB of excursion and frequency of 120 Hz in signals that traverses the power network, especially in industrial locations. This periodical fading is equivalent to the classical working principle of an amplitude modulator, whose output signal is

$$y(t) = x(t)\cos(2\pi f_c t), \tag{73}$$

for the case of double-side band modulation and where f_c represents the carrier frecuency. These systems are often presented as examples of non LTI systems, because, in fact, they are LPTV systems with period $T_c = 1/f_c$ and impulse response

$$h(t,u) = \delta(t-u)\cos(2\pi f_c t). \tag{74}$$

It can be observed that the rotational phasor of the carrier yields the time variation, in the PLC case it is related to the mains frequency $f_c = f_0 = \frac{1}{T_0}$ (or often $f_0 = \frac{2}{T_0}$ as in [16]). The system frequency response is the FT of $h(t,u)$ in variable u

$$H(t,f) = \cos(2\pi f_0 t), \tag{75}$$

which is independent of f, i.e. flat (distortion-less), but sinusoidally varying in time. FS coefficients of the frequency response can be easily identified as

$$H^{\alpha}(f) = \begin{cases} 1/2, & \textit{if } \alpha \in \{-1, 1\} \\ 0, & \text{otherwise.} \end{cases} \tag{76}$$

When a stationary signal, with a PSD $S_X(f)$, excites this modulating system, the output signal is CS and has the cyclic spectra (see (60))

$$\begin{aligned} S_Y^{\alpha}(f) = &\frac{1}{2}H^{\alpha+1}\left(f - \frac{\alpha+1}{T_0}\right) S_X\left(f - \frac{\alpha+1}{T_0}\right) \\ &+ \frac{1}{2}H^{\alpha-1}\left(f - \frac{\alpha-1}{T_0}\right) S_X\left(f - \frac{\alpha-1}{T_0}\right). \end{aligned} \tag{77}$$

Being the first one, which corresponds to $\alpha = 0$, the PSD of the output signal:

$$S_Y^0(f) = \frac{1}{4}S_X\left(f - \frac{1}{T_0}\right) + \frac{1}{4}S_X\left(f + \frac{1}{T_0}\right). \tag{78}$$

This a well-known results that highlights the spectral translation inherent to linear modulation, useful for heterodynation. Finally, let us calculate the output signal mean, which is obviously periodic:

$$m_Y = m_X H(t,0) = m_X\cos(2\pi f_0 t). \tag{79}$$

This power line modulation feature also appears when an electrical load (e.g. an appliance) generates noise and it is introduced to the power network, even if such noise is stationary it can changes to CS at the receiver after traversing the LPTV channel.

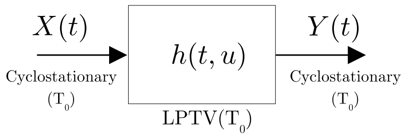

FIGURE 10. LPTV filtering of a CS, both with the same period T_0.

C. FILTERING CYCLOSTATIONARY SIGNALS THROUGH LPTV SYSTEMS

In this section, the general case of getting an CS signal at the output of a LPTV system that is excited by a CS signal is addressed as depicted in Fig.10. Two different situations appear: when both the input signal period and the system period are the same, and when they are not. The analysis here is focused only in the former case, which is the case adequate to study PLC signals and systems. For a more extensive treatment, covering the second situation, the works by W.A. Gardner can be consulted [6], [10].

Property 6: When filtering a CS signal $X(t)$ with a LPTV system, the relation between output and input mean cyclic coefficients is given by

$$m_Y^{\alpha} = \sum_{\beta=-\infty}^{+\infty} m_X^{\alpha-\beta} H^{\beta}\left(\frac{\alpha-\beta}{T_0}\right). \tag{80}$$

Proof: $X(t)$ and the output signal $Y(t)$ are related by means of the superposition integral with LPTV system response $h(t,u)$ in (59)

$$\begin{aligned} E[Y(t)] &= E[\int_{-\infty}^{+\infty} h(t, t-\tau)X(t-\tau)\mathrm{d}\tau] \\ &= \int_{-\infty}^{+\infty} h(t, t-u)m_X(t-\tau)\mathrm{d}\tau. \end{aligned} \tag{81}$$

By expanding in FS the input mean and after applying the FT in variable τ, the following expression can be derived:

$$\begin{aligned} m_Y(t) &= \int_{-\infty}^{+\infty} h(t, t-\tau) \sum_{\gamma=-\infty}^{+\infty} m_X^{\gamma} e^{j2\pi\gamma(t-\tau)/T_0}\mathrm{d}\tau \\ &= \sum_{\gamma=-\infty}^{+\infty} m_X^{\gamma} e^{j2\pi\gamma t/T_0} \int_{-\infty}^{+\infty} h(t, t-\tau)e^{-j2\pi\gamma\tau/T_0}\mathrm{d}\tau \\ &= \sum_{\gamma=-\infty}^{+\infty} m_X^{\gamma} H\left(t, \frac{\gamma}{T_0}\right) e^{j2\pi\gamma t/T_0}. \end{aligned} \tag{82}$$

At this point it is convenient to expand the periodic response $H(t, \gamma/T_0)$, defined in (55),

$$\begin{aligned} m_Y(t) &= \sum_{\gamma=-\infty}^{+\infty} m_X^{\gamma} \sum_{\beta=-\infty}^{+\infty} H^{\beta}\left(\frac{\gamma}{T_0}\right) e^{j2\pi\beta t/T_0} e^{j2\pi\gamma t/T_0} \\ &= \sum_{\alpha=-\infty}^{+\infty} \sum_{\beta=-\infty}^{+\infty} m_X^{\alpha-\beta} H^{\beta}\left(\frac{\alpha-\beta}{T_0}\right) e^{j2\pi\alpha t/T_0}, \end{aligned} \tag{83}$$

where a change of sum indexes has been made to identify terms in the FS expansion of the periodic mean (in (4)) to derive (80), concluding the proof. ■

This expression is a formal generalization of the previous ones in this paper. By setting $\beta = 0$ in it, the mean coefficients after a LTI filtering of a CS signal case, in (21), are reached. Alternatively, by setting $\alpha = \beta$ in it, the mean coefficients after a LPTV filtering of a CS signal case, in (61), are given. When both restrictions hold, i.e. $\alpha = \beta = 0$, the well-known relation for the output mean after a LTI filtering of stationary signals, $m_Y = m_X H(0)$, is addressed.

Property 7: When filtering a CS signal $X(t)$ with a LPTV system, the output signal cyclic spectra are related to $X(t)$ cyclic spectra by means of the expression in (84).

$$S_Y^{\alpha}(f) = \sum_{\beta=-\infty}^{+\infty}\sum_{\gamma=-\infty}^{+\infty}\left[H^{\beta}\left(f-\frac{\alpha+\beta}{T_0}\right)\right]^* \times H^{\gamma}\left(f-\frac{\gamma}{T_0}\right)S_X^{\alpha+\beta-\gamma}\left(f-\frac{\gamma}{T_0}\right) \quad (84)$$

Proof: By starting from the autocorrelation definition in (64) and after substituting (65) and (66), it yields

$$\begin{aligned} R_Y(t,t+u) &= \iint_{-\infty}^{+\infty}\sum_{\beta=-\infty}^{+\infty}\sum_{\gamma=-\infty}^{+\infty} e^{j2\pi(\gamma-\beta)t/T_0}e^{j2\pi\gamma u/T_0} \\ &\quad \cdot[h^{\beta}(s)]^*h^{\gamma}(v)E[X^*(t-s)X(t+u-v)]\mathrm{d}s\mathrm{d}v \\ &= \sum_{\beta=-\infty}^{+\infty}\sum_{\gamma=-\infty}^{+\infty} e^{j2\pi(\gamma-\beta)t/T_0}e^{j2\pi\gamma u/T_0} \\ &\quad \cdot\iint_{-\infty}^{+\infty}[h^{\beta}(s)]^*h^{\gamma}(v)R_X(t-s,t+u-v)\mathrm{d}s\mathrm{d}v, \end{aligned} \quad (85)$$

and the input autocorrelation term, which is periodic, can be expanded in a FS (7)

$$R_X(t-s,t+u-v) = \sum_{\delta=-\infty}^{+\infty} R_X^{\delta}(u+s-v)e^{j2\pi\delta(t-s)/T_0} \quad (86)$$

to include it in (85) and reach to

$$\begin{aligned} R_Y(t,t+u) &= \sum_{\beta=-\infty}^{+\infty}\sum_{\gamma=-\infty}^{+\infty}\sum_{\delta=-\infty}^{+\infty} e^{j2\pi(\gamma-\beta+\delta)t/T_0}e^{j2\pi\gamma u/T_0} \\ &\quad \cdot\iint_{-\infty}^{+\infty}[h^{\beta}(s)]^*h^{\gamma}(v)R_X^{\delta}(u+s-v)e^{-j2\pi\delta s/T_0}\mathrm{d}s\mathrm{d}v \\ &= \sum_{\beta=-\infty}^{+\infty}\sum_{\gamma=-\infty}^{+\infty}\sum_{\delta=-\infty}^{+\infty} e^{j2\pi(\gamma-\beta+\delta)t/T_0}e^{j2\pi\gamma u/T_0} \\ &\quad \cdot[h_1^{\beta,\delta}(-u)*h^{\gamma}(u)*R_X^{\delta}(u)]. \end{aligned} \quad (87)$$

For this latter step, convolution properties (similarly to derivation of (69)) have been employed together with the ancillary function

$$h_1^{\beta,\delta}(t) = [h^{\beta}(t)e^{j2\pi\delta t/T_0}]^*. \quad (88)$$

Now, a change of index is made $\alpha = \gamma - \beta + \delta$, to give (87) the form of a FS:

$$\begin{aligned} R_Y(t,t+u) &= \sum_{\alpha=-\infty}^{+\infty}\sum_{\beta=-\infty}^{+\infty}\sum_{\gamma=-\infty}^{+\infty} e^{j2\pi\gamma u/T_0} \\ &\quad \cdot[h_1^{\beta,\delta}(-u)*h^{\gamma}(u)*R_X^{\alpha+\beta-\gamma}(u)]e^{j2\pi\alpha t/T_0}, \end{aligned} \quad (89)$$

which supports that the output signal is CS with period T_0, since both its mean and autocorrelation are periodic. From (89) the input-output relation between cyclic coefficients of the autocorrelation can be identified

$$R_Y^{\alpha}(u) = \sum_{\beta=-\infty}^{+\infty}\sum_{\gamma=-\infty}^{+\infty} e^{j2\pi\gamma u/T_0} \cdot[h_1^{\beta,\delta}(-u)*h^{\gamma}(u)*R_X^{\alpha+\beta-\gamma}(u)], \quad (90)$$

which constitutes a valuable intermediate result.

By calculating the FT of the last expression, given by (14), the output signal cyclic spectra are obtained:

$$S_Y^{\alpha}(f) = \sum_{\beta=-\infty}^{+\infty}\sum_{\gamma=-\infty}^{+\infty} H_1^{\beta,\delta}\left(-f+\frac{\gamma}{T_0}\right) \cdot H^{\gamma}\left(f-\frac{\gamma}{T_0}\right)S_X^{\alpha+\beta-\gamma}\left(f-\frac{\gamma}{T_0}\right). \quad (91)$$

A further simplification is possible by substituting in (90) the FT of the auxiliary function $h_1^{\beta,\delta}(u)$,

$$\begin{aligned} H_1^{\beta,\delta}(f) &= \left[H^{\beta}\left(-f-\frac{\delta}{T_0}\right)\right]^* \\ &= \left[H^{\beta}\left(-f-\frac{\alpha+\beta-\gamma}{T_0}\right)\right]^*, \end{aligned} \quad (92)$$

to reach to the desired expression in (84), which concludes the proof. ■

This relation represents a formal generalization of previous cases presented in this work. First, by setting in it $\beta = \gamma = 0$ the cyclic spectra of a CS signal after a LTI filtering case, (22), is reached. By setting $\gamma = \alpha + \beta$ the output cyclic spectra after LPTV filtering of a stationary signal case, (60), is given. When both restrictions hold, i.e. $\alpha = \beta = \gamma = 0$, the well-known expression for PSD of stationary signals LTI filtering, $S_Y(f) = S_X(f)|H(f)|^2$, is addressed.

A practical example: Wiener Filtering. In signal estimation theory, it is well known that the optimal linear estimator of a stationary random process, under a minimum mean square error (MMSE) criterion, is the Wiener filter. In the case of a CS signal, such estimator is the cyclic Wiener filter [36]. This has been studied for PLC applications in [37], where two estimators of this kind, realized by means of an adaptive LPTV filter with the structure of a FRESH filter like depicted in Fig.7, are used. First one is designed to estimate the CS noise at the receiver and try to cancel it. Second one is employed to estimate the received OFDM signal that is contaminated by the non-canceled CS noise.

V. A PRACTICAL SIMPLIFICATION FOR PLC CHANNELS: SLOW VARIATION LPTV MODEL

PLC channels can be modeled as LPTV systems but there are special features that simplify their study [17], [21]. The most important one is the long period T_0 compared with the channel impulse response duration, which results in a quite slow variation of the system that causes a low Doppler spread in the signals. Consequently, the channel coherence time (inversely related to the Doppler spread) is much longer than the impulse response duration or the delay spread and the channel is referred to as underspread. Measurements in [21], reveals that for intervals of approximately 20 μs the channel variation is not relevant, i.e. three orders of magnitude with respect to the mains cycle. Even more, estimations of a conservative upper-bound for the channel Doppler spread are below 2 kHz, what leads to a channel coherence time greater than 600 μs. Under this circumstances, some simplification of the relations presented in previous sections are possible.

1) FILTERING OF DETERMINISTIC SIGNALS

The output of a LPTV system is described by expressions in (50) and (51) where, assuming that excitation signal duration is much shorter than the system period T_0, the exponential terms have a slow variation and can be considered as invariant (since they are harmonics of $1/T_0$). These conditions are reasonable for packet oriented transmission in PLC systems and, hence, (51) simplifies to

$$y(t) \simeq y_\sigma(t) = \sum_{\alpha=-\infty}^{+\infty} y_\sigma^\alpha(t) e^{j2\pi\alpha\sigma/T_0}. \quad (93)$$

From which, it can be demonstrated (see [21]-Appendix I.B.1) that the following expression holds

$$Y_\sigma(f) \simeq H(t,f)|_{t=\sigma} \cdot X_\sigma(f), \quad (94)$$

that represents an input-output relation based on the definition of a short-term approximation restricted to a time interval $t \approx \sigma$ (defined as $t \in [\sigma - \Delta t, \sigma + \Delta t]$ with a Δt smaller enough), of the output and input signals FT and is a simplification of (57) and (58).

This analysis can be seen as a slow-variation decomposition of the LTPV in the form of a series of successive and periodical LTI systems synchronized with the mains period; and the input signal faces one of these, depending on the time phase at which it arrives [17].

2) FILTERING OF RANDOM SIGNALS

Regarding the filtering of random signals, similar considerations can be assumed that yield $H(f)$ and $S_x(f)$ to be essentially non selective to a frequency-shift of significant harmonics of the fundamental frequency (in the sense that higher harmonics can be disregarded under the slow-variation approximation). This way, a simplified form of (84) can be expressed as (see [21]-Appendix I.B.2)

$$\begin{aligned} S_Y^\alpha(f) &\simeq \sum_{\beta=-\infty}^{+\infty} [H^\beta(f)]^* \cdot [H^{\alpha+\beta}(f) \odot S_X^{\alpha+\beta}(f)] \\ &= [H^{-\alpha}(f)]^* \odot H^\alpha(f) \odot S_X^\alpha(f); \end{aligned} \quad (95)$$

where the symbol $\odot$ denotes a convolution sum in the cyclic coefficients domain. For the validity of (95), the order of significant coefficients of $H^\alpha(f)$ must be low e.g. harmonics of the cyclic response of very high order can be neglected. These assumptions are reasonable for real PLC channels, because are counterparts in the frequency domain of the slow variation in time. From the analysis of actual channel measurements in [17], it is observed that no more than 30 coefficients of $H^\alpha(f)$ are significant in the low frequency band (under 2 MHz) and 15 for higher bands. Considering significant those above -40dB with respect to the maximum value.

From (95) an interesting result can be derived to obtain the output signal IPSD in the following form (see [21]-Appendix I.B.2):

$$S_Y(t,f) \simeq |H(t,f)|^2 \cdot S_X(t,f). \quad (96)$$

This relation shows a similar structure to the well-known expression of the PSD of a stationary signal filtered by an LTI system, but applied here to the IPSD after LPTV filtering of CS input signals.

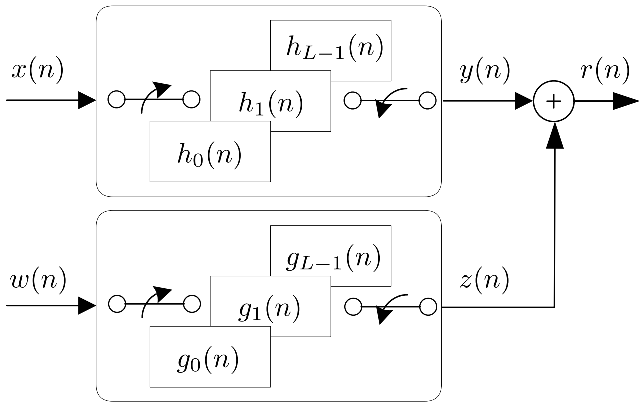

FIGURE 11. Scheme of the slow variation cyclic channel model implementation: A bank of cyclically commuted LTI filters is used to represent both the channel response and the noise.

This slow variation approach is illustrated in the scheme of Fig.11, which shows the expansion of a LPTV discrete-time system in a set of states (for each invariance interval), characterized with LTI filters of responses $h_l(n)$, $l \in \{0, L-1\}$, selected in a synchronized manner with mains cycle. L is the cardinal of the set of channel states. The additive CS noise is modeled in an analogous manner with the bank of filters of responses $g_l(n)$ excited with stationary noise. This feature of PLC channels has served to carry out measurements, to develop models of reasonable complexity and to implement modem prototypes, of which example are discussed in the next section.

VI. STUDIES ABOUT THE CYCLIC BEHAVIOR OF PLC CHANNELS

The cyclic nature of power line appliances and its impact on the system response and received noise have been extensively studied in the last two decades. A selection of the relevant works is here reviewed in a somewhat chronological order and classified in two groups: those devoted mainly to the channel time variation and those focused on CS noise.

A. LPTV CHANNEL MODELING

After the inspiring work in [16], where the cyclic fading of PLC channels was pointed out, many papers devoted to LPTV models for these channels have appeared. As mentioned in section II, a characterization of PLC cyclic short-time variation was shown in [17]–[19], [21]. It was based on the analysis of devices and channel measurements with signal processing techniques based on the slow variation LPTV model, with a cyclic sampling of the channel. Actually, in [38], a channel simulator assuming this slow variation can be found, which implements an interpolation between LTI channel states (zero-order hold or linear). An alternative LPTV channel model is proposed in [22], also under a slow-variation approach, by using a discrete-time block model with wavelets. Another example of a model implementation, following this approach, by using cascaded IIR (infinite impulsive response) filters was given in [39].

A proposal of LPTV channel adequate for broadband indoor PLC is described in [40], based on a simple topology of transmission lines with three loads. The per-unit length line parameters are taken from typical cables, whose line sections and lengths are random variables with common probability distributions. The load impedance functions are frequency selective, calculated from physical parameters (also modeled with random variables), and cyclic time-varying by following the slow-variation approach.[8]

Later, in [23], the PLC channel modeling is revisited by proposing a Zadeh's expansion of the LPTV channel that is equivalently represented as a bank of LTI channels as the one shown in Fig.7 with the so-called FRESH approach. The set of LTI responses are defined with a simple two-tap model, whose random coefficients are generated according to a statistical fit to actual measurements of channel delay spread and mean attenuation. A discrete-time implementation following this FRESH modeling can be also found in [41], [42] where a method to perform a frequency domain estimation of the channel cyclic response filters (56), is proposed. The estimation is validated with measurements of PLC channels up to 1 MHz and a number of 11 real-valued filters (corresponding to the fundamental plus 5 harmonics of the mains frequency) is found adequate to represent the channel time variation.

In [20], [26], a model is presented to estimate bounds of both the time and frequency domain responses of the PLC channel due to the presence of loads time variation.

B. CYCLOSTATIONARY NOISE CHARACTERIZATION

A pioneer attempt to model PLC cyclic noise can be found in [43], where noise registers of the three-phase network at a substation are studied. A harmonic model of periodical impulses is proposed, with six components in a period of the mains voltage (attributed to the positive and negative transitions of switching devices at each phase circuit). Also, in [44], the inter-arrival time between noise impulses is investigated on noise measurements at indoor power lines and a periodicity with half the mains cycles clearly arose.

A more elaborated model for NB-PLC noise up to 500 kHz was proposed in [45], which is supported by the observation of a clear periodicity in the autocorrelation function of noise signals measured at indoor power networks. An interesting feature pointed out in this work is that, when the noise is sampled periodically with half the mains period and synchronously at different phases of the period, the amplitude at each phase can be considered as almost Gaussian. However, with no synchronization with the mains cycle, the noise seems impulsive. Hence, the noise model is the sum of contributions from different sources, each of them characterized as a CS Gaussian process whose samples have a variance that is a periodical function with half the mains period. The power spectral density of this components is assumed to decrease with frequency exponentially. Parameters of the model are adjusted according to actual measurement results.

These authors went beyond in the exploration of the cyclic properties of PLC noise in [46] by analyzing longer records of measurements in the time and frequency domain to extract a better statistical characterization of it cyclostationarity. As a result, they proposed an enhanced mathematical noise model in [47], which can represent its non-stationary and non-white features by probability density functions with a small set of parameters. A procedure to generate noise waveform according to the model is also provided. An evolution of the preceding model is available in [48], where the authors make use of the slow-variation approach of LPTV described in section V to filter Gaussian random signals to obtain the CS noise. The number of noise states in the mains period is estimated by applying thresholds to spectrograms of the IPDS given by a discrete-time formulation of (11).

The authors of [49] presented an exhaustive analysis and classification of PLC noise for indoor PLC broadband systems, with emphasis in the impulsive part and its cyclic properties and frequency content. Alternatively, the work in [50] is focused on narrowband systems for smart-grids applications in the 3-500 kHz band. In it, results of simple models of CS noise based on a Gaussian process with a time-varying variance are reviewed. In [31], the CS noise observed in a NB-PLC system operating in a band from 3 to 148.5 kHz

[8] A software channel generator based on the implementation of such model is available at www.plc.uma.es.

is analyzed in several urban locations, exhibiting variations of more than 20 dB along mains cycle. Another extensive study of NB-PLC noise measured in the CENELEC-A band (from 3 to 95 kHz) at 106 locations, in rural, semi-urban and urban areas, for multiple-input multiple-output (MIMO)[9] 3x3 systems is given in [51] (i.e. leading to 954 streams). It can be observed that 50% of the measured noise registers exhibit a peak energy variation along the mains cycle higher than 13 dB.

An alternative approach to model CS noise in PLC is provided in [52], were a LPTV FRESH filtering as depicted in Fig.7 is considered to synthesized NB-PLC noise samples from stationary noise. Additionally, the idea is extended to cover MIMO NB-PLC noise modeling. Recent works have studied CS noise classification in MIMO PLC according to classical models [53].

VII. PLC SYSTEM DESIGN WITH CYCLIC-AWARE STRATEGIES

We summarize here some works that illustrate the importance that cyclic signals and systems have on PLC equipment design. First, it is worth mentioning some studies where signal processing algorithms at receivers are designed by considering the cyclic features of the channel and noise. All of them are part of OFDM systems, which constitute the dominant technology in this context.

There is a natural consequence of the channel time variation, which is studied in [35]: the distortion in form of ICI caused by the spectral broadening expressed in (52). This establishes a theoretical bound in the OFDM system performance and has the practical consequence of limiting the number of carriers, beyond which this ICI dominates against the distortion due to the channel frequency selectivity.

It is observed in measurements is studied for PLC multicarrier systems and the achievable performance with an optimized setting of the system parameters is evaluated.

In particular, several of them are devoted to *channel estimation and equalization*. For instance, in [54], the design of a prototype on FPGA (field-programmable gate array) of a PLC multi-carrier modem is presented. It incorporates a FEQ (frequency domain equalizer) with an adaptive filter, based on the LMS (least mean squares) algorithm, to track the channel time variation. The work in [55] investigates pilot based channel estimation for broadband systems and proposes a linear arrangement of the pilots distributed with a slope in time and frequency to reduce interpolation errors. Also in [37], an adaptive filtering scheme for wide-sense CS signals is described, and simulations of practical applications to channel estimation in NB-PLC scenarios are presented.

Many works deal with *bit-rate adaptation* algorithms adapted to exploit the cyclic channel characteristics. In [29], [56], an LPTV-aware bit loading of the OFDM carriers is proposed and the obtained performance improvement evaluated. They take into account a partition of the mains period in smaller slots of almost time-invariant regions, by following the slow-variation approach. SNR gains are reported of about 3 dB for a BER of 10^{-3}, when using cyclic channel adaptation with a constellations set of BPSK and M-QAM ($M \in \{4, 16, 64, 256, 1024\}$) with respect to a fixed parameter setting case. Similar works can be found in [57]–[59], for bit and power allocation strategies, where the channel simulator proposed in [40] are used to test the algorithms performance, exhibiting SNR gains of about 4-5 dB. A cyclic bit loading method for a PLC system in the CENELEC A-band is evaluated in [60]. It divides the mains period into 8 time slots and employs a tone mapping between 0 and 6 bits per M-QAM symbol. The method is tested with a set of 1659 actual measured channels and improvements of the attainable bit-rate exceeding 30% are reported for certain configurations. Similar strategies for bit and power allocation in the time-frequency domain were studied in [61], for indoor NB-PLC. Results over channel simulations doubling the data rate when adapting the modulation M-QAM ($M \in \{0, 4, 16, 64\}$) and a BER constraint of $10^{-4}$are presented. Finally, in [62], strategies for bit loading both in time and frequency are studied with a trade-off between computational complexity and bit rate gain. The algorithms are validated with a measurement campaign carried out in houses and apartments, with more than 7000 PLC channels.

There are other works related to *synchronization* issues, like the one in [63], where multicarrier timing recovery schemes for systems over indoor power lines are analyzed. The influence of the channel cyclic short-term variations is assessed and improvements are proposed: a new phase error estimator and the introduction of notch filters in the timing recovery loop at harmonics of the mains frequency. In [64], a pilot-aided joint estimation of both the carrier frequency offset and the channel impulse response for LPTV channels, applied to NB-PLC systems up to 500 kHz, is discussed.

In the field of *noise cancellation* strategies, some papers can be mentioned. The early work in [43] suggested a noise canceling strategy for PSK transmission by means of an LMS adaptive filtering for CS noise in the form of impulsive periodical waveforms. Later, in [65], two strategies were discussed to improve the robustness of OFDM systems to periodical noise impulses synchronized with the mains period: to extend the OFDM symbol and to use a linear predictive filter in the frequency domain to cancel it. In [66], receiver strategies for CS noise reduction by using the FRESH filter approach are explained, they are applied to low SNR NB-PLC systems that implement OFDM. In [67], an improvement of the previous technique for CS noise mitigation designed with estimators also based on the FRESH filtering is discussed, with focus on PLC receivers of systems with single-input multiple-output.

Regarding *Medium Access Control* (MAC) strategies, in [68], a cross-layer protocol named Opportunistic CSMA (Carrier-Sense Multiple-Access) is proposed. It consist in a modification of the conventional contention-based CSMA

[9]In PLC, MIMO systems are established over the three wires (phase, neutral and earth).

with Collision Avoidance (CSMA/CA) protocol, taking into account CS noise in home PLC networks. Time periods, or SNR regions, in a mains cycle are determined depending on the periodical SNR values compared with some thresholds and different access policies are defined for such periods, according to the quality of service demanded by the users.

Finally, there are many *PLC standards* where considerations on cyclic signals and systems have a primary role. In the IEEE Standard 1901 ''for Broadband over Power Line Networks: Medium Access Control and Physical Layer Specifications'' [69], mains cycle synchronization is obliged for all devices, for instance the medium access control allows a time-division multiplexing access (TDMA) scheme synchronous with mains cycle to give stability to the TDMA allocations. A dynamic channel adaptation procedure is defined so that the employed modulation, referred as tone maps, can follow the channel periodical signal to noise ratio (the so-called AC line cycle adapted tone map). The coexistence mechanism between equipment of different vendors requires the mains cycle synchronization as well. Similar procedures for channel adaptation, under a MAC frame synchronized with mains cycle, are defined in ITU-T Recommendation 6.9960 ''Unified high-speed wire-line based home networking transceivers - System architecture and physical layer specification'' [70], also devoted to broadband PLC. They are called bit allocation tables (BAT) that, after adequate channel estimation procedures, may be defined that are valid for only specific portions of the MAC cycle (BAT region).

Regarding NB-PLC regulation, the IEEE 1901.2 standard [34], includes an annex that describes cyclic aspects of the channel. CS noise observed in actual power lines for smart-grid communications is addressed, measurements of spectrograms synchronized with mains period are shown and simplified models to simulate periodical impulsive CS noise are presented. Also measurements of loads time variation and implications on system periodical response are given there.

VIII. CONCLUSION

This work deals with the cyclic variation observed in many PLC channels caused by the presence of mains voltage. Theoretical results of the interplay between CS random signals and LPTV systems for the particular case of sharing the same period are presented and justified. The focus is on their application to PLC channels, what simplifies the general relations for LPTV filtering of random or deterministic signals that are rather complicated. Proofs for certain properties that can be addressed with a reasonable mathematical treatment are given. The implications of the slow variation character of actual PLC channels in the LPTV model are presented and helpful approximations for the design of PLC systems are provided. A review of works devoted to the study of PLC channels cyclic behavior is provided, as well as a discussion about how current PLC technology exploits this periodical characteristics, e.g. by means of adaptive modulation.

Despite the significant advances experienced by PLC modems in the last decades, there is still room for improvement by incorporating additional refinements related to these cyclic properties, which have a higher computational complexity but that could be affordable with the continuous development of signal processing hardware. For instance, as future research lines can be mentioned: the characterization of the cyclic relation between noise signals in multiple-output receivers and the design of canceling algorithms that exploit this relation (extending the ideas in [71], [72]). A promising development is the implementation of in-band full-duplex PLC systems that can benefit from these features, for instance with echo cancelers that take into account the cyclic behavior of the echo channels [73]. Also, more theoretical works to explore these cyclic channels capacity like the one in [74] are interesting.

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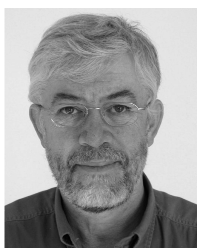

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