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A Novel Method for Estimation of The Fundamental Parameters of Distorted Single Phase Signals

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Abstract

This paper proposes a new method for parameter estimation of distorted single phase signals, through an improved demodulation-based phase tracking incorporated with a frequency adaptation mechanism. The simulation results demonstrate the superiority of the proposed method compared to the conventional SOGI (Second-Order Generalized Integrator)-based approach, in spite of the dc-offset and harmonic distortions.

Keywords: Phase-Locked Loop (PLL); Frequency Estimation; Amplitude Estimation; Second-Order Generalized Integrator (SOGI); Disturbance; DC-offset.

1. Introduction

The fundamental parameters of signals have to be estimated in many important fields, such as power system analysis, industrial electronics, stability analysis and communications [1]–[4]. The estimation methods are generally classified into frequency and time domains. In the real-time applications, the time domain methods are faster and more accurate than the other one [5]. The methods based on the phase-locked loop (PLL) or frequency-locked loop (FLL) are the conventional time domain approaches that are able to estimate important parameters. A number of PLL structures are applied only for sinusoidal signals such as SRF-PLL(synchronous reference frame-based PLL) [6]; but the faults and disturbances in the input signal may lead to parameter estimation error. Consequently, estimation of the fundamental parameters in distorted conditions (such as for power quality and protection applications) requires robustness against disturbances such as harmonics, dc-offset, sag, swell and frequency deviations. Therefore, an important issue here is the robust estimation of the amplitude, phase angle and frequency of distorted single phase signals.

Some structures have been proposed for providing this requirements each of which can estimate only some of the parameters for single phase applications. For example VSPF-PLL can estimate phase and frequency; however, it cannot estimate the amplitude without a pre-estimation of the amplitude [7]. Another structure is CDSC-PLL that similar to LPN-PLL is not robust against dc-offset and frequency changes [1], [5].

Recently, a frequency locked loop (SOGI-FLL) has been developed to generate the quadratic signals for estimation of the phase and amplitude in addition to the frequency [8], [9]. This structure can be modified to estimate the dc-offset to improve its performance [10] and to facilitate low frequency electromagnetic oscillation monitoring [2]; nevertheless its performance is limited against the harmonic and dc-offset distortions, as it is demonstrated in simulation section.

Our purpose herein is presentation of a novel method for estimation of all important fundamental parameters of distorted single phase signals. The structure of the demodulation-based PLL (such as LPN-PLL) is improved to achieve this purpose. The proposed modification takes the advantages of accurate and fast response of the demodulation-based PLLs and their flexibility against filter types, and provides a superior performance over SOGI-FLL. The proposed method has two dynamic types caused by tuning its low pass filters. This method has no steady state error against the harmonic distortion, provides more robustness against the dc-offset and a better amplitude estimation.

The rest of the paper is organized as follows. Section II introduces the conventional demodulation-based PLL. The proposed method is described in Section III. The effectiveness of the proposed method is demonstrated by simulation results in Section IV. Finally, the conclusion is presented in Section V.

1. Conventional Demodulation-based PLL

The structure of the conventional demodulation-based PLL is shown in Fig. 1. This structure takes a feed-forward angular frequency to generate a sinusoidal signal that has the same frequency to extract two low frequency sinusoidal components for calculating θ_0.

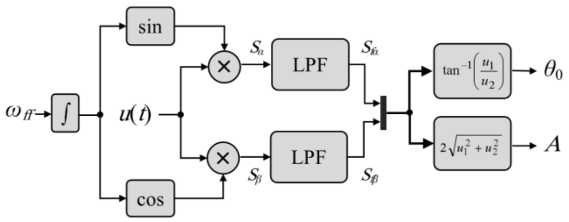

Fig. 1. Structure of the conventional demodulation-based PLL.

At the first stage, the input signal is multiplied by a primary generated signal, given as

$$u(t) = A\sin(\omega t + \theta_0) + A_{\mathrm{dc}}, \tag{1}$$

where A, ω, θ0 and Adc are respectively the amplitude of sinusoidal component, angular frequency, initial phase angle and the value of dc component. The following equations represent the demodulated signals [1].

Sα and Sβ have two terms for any harmonic, including low frequency and high frequency components. Moreover, a sinusoidal component having the fundamental frequency f, appears due to the dc component. These components, except the low frequency component of the first harmonic, are then eliminated by two low pass filters (LPFs). If the input signal is free of the dc component and the even harmonics, the oscillation frequencies are even coefficients of the fundamental frequency. Therefore, we can tune the filters to eliminate the oscillations having frequencies of 2f, 4f and higher even frequencies; otherwise, one has to tune the filter for all frequencies. Having the precise filters and feed-forward frequency, the filter outputs are as follows.

Subsequently, the orthogonal filter outputs are used to estimate the amplitude and initial phase angle.

$$\begin{aligned} S_\alpha &= u(t)\times\sin(\omega_{ff}t) \\ &= A\sin(\omega t+\theta_0)\times\sin(\omega_{ff}t) \\ &\quad+\sin(\omega_{ff}t)\sum_h A_h\sin(h\omega t+\theta_{0,h}) \\ &\quad+A_{\mathrm{dc}}\sin(\omega_{ff}t) \\ &= \frac{1}{2}A\begin{bmatrix}\cos\big((\omega-\omega_{ff})t+\theta_0\big)- \\ \cos\big((\omega+\omega_{ff})t+\theta_0\big)\end{bmatrix} \\ &\quad+\frac{1}{2}\sum_h\begin{bmatrix}\cos\big((h\omega-\omega_{ff})t+\theta_{0h}\big)- \\ \cos\big((h\omega+\omega_{ff})t+\theta_0\big)\end{bmatrix} \\ &\quad+A_{\mathrm{dc}}\sin(\omega_{ff}t), \end{aligned} \tag{2}$$

$$\begin{aligned} S_\alpha &= u(t)\times\sin(\omega_{ff}t) \\ &= A\sin(\omega t+\theta_0)\times\sin(\omega_{ff}t) \\ &\quad+\sin(\omega_{ff}t)\sum_h A_h\sin(h\omega t+\theta_{0,h}) \\ &\quad+A_{\mathrm{dc}}\sin(\omega_{ff}t) \\ &= \frac{1}{2}A\begin{bmatrix}\cos\big((\omega-\omega_{ff})t+\theta_0\big)- \\ \cos\big((\omega+\omega_{ff})t+\theta_0\big)\end{bmatrix} \\ &\quad+\frac{1}{2}\sum_h\begin{bmatrix}\cos\big((h\omega-\omega_{ff})t+\theta_{0h}\big)- \\ \cos\big((h\omega+\omega_{ff})t+\theta_0\big)\end{bmatrix} \\ &\quad+A_{\mathrm{dc}}\sin(\omega_{ff}t), \end{aligned} \tag{3}$$

$$\begin{aligned} S_\beta &= u(t)\times\cos(\omega_{ff}t) \\ &\quad A\sin(\omega t+\theta_0)\times\cos(\omega_{ff}t) \\ &\quad+\cos(\omega_{ff}t)\sum_h A_h\sin(h\omega t+\theta_{0,h}) \\ &\quad+A_{\mathrm{dc}}\cos(\omega_{ff}t) \\ &= \frac{1}{2}A\begin{bmatrix}\sin\big((\omega-\omega_{ff})t+\theta_0\big)+ \\ \sin\big((\omega+\omega_{ff})t+\theta_0\big)\end{bmatrix} \\ &\quad+\frac{1}{2}\sum_h\begin{bmatrix}\sin\big((h\omega-\omega_{ff})t+\theta_{0h}\big)+ \\ \sin\big((h\omega+\omega_{ff})t+\theta_0\big)\end{bmatrix} \\ &\quad+A_{\mathrm{dc}}\cos(\omega_{ff}t). \end{aligned} \tag{4}$$

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$$S_{f\alpha} = \frac{1}{2}A\cos\left((\omega-\omega_{ff})t+\theta_0\right) = \frac{1}{2}A\cos(\theta_0), \quad \forall\omega=\omega_{ff}$$

$$S_{f\beta} = \frac{1}{2}A\sin\left((\omega-\omega_{ff})t+\theta_0\right) = \frac{1}{2}A\sin(\theta_0). \quad \forall\omega=\omega_{ff} \tag{5}$$

The conventional low pass filters have to be of high-order to have a sharp behavior and a low cut-off frequency. An approach was proposed in [1] to eliminate oscillations by a fourth-order low pass notch filter, so that the cut-off frequency is tuned to locate the notch at the second-order component; but, this approach does not cover the neighbor components (such as 1st and 3th components caused by the input dc-offset and even harmonics). Another drawback to the conventional demodulation-based PLL is using feed-forward frequency that needs a separate frequency estimation structure and may cause a steady state error. Therefore, the conventional structure should be improved to estimate the fundamental parameters in distorted conditions.

Moving average filter (MAF) is a finite impulse response (FIR) filter. MAF response for input x(τ) is formulated as follows [11]:

$$\bar{x}(t) = \frac{1}{T_w}\int_{t-T_w}^{t} x(\tau)\mathrm{d}\tau, \tag{6}$$

where Tw is the window length in second. This filter calculates the average of input signal from $t-T_w$ to t. MAF is practically implemented in discrete form as follows:

$$\bar{X} = \frac{1}{N}\sum_{n=0}^{N-1} x(-n), \tag{7}$$

x(0) is the last input sample and x(−N+1) is N-th previous sample, where $N=\mathrm{round}(T_w/T_s)$ is the sampling window length with Ts as the sampling time. The calculation error reduces by interpolation and increasing the sampling frequency [11].

The transfer function of MAF is given as (8) and Its frequency response is represented in (9) and illustrated in Fig. 2.

$$G_{MAF}(s) = \frac{1-e^{-sT_w}}{sT_w}. \tag{8}$$

$$G_{MAF}(j\omega) = \frac{2\left|\sin\left(\frac{\omega T_w}{2}\right)\right|}{\omega T_w} \angle -\frac{\omega T_w}{2}. \tag{9}$$

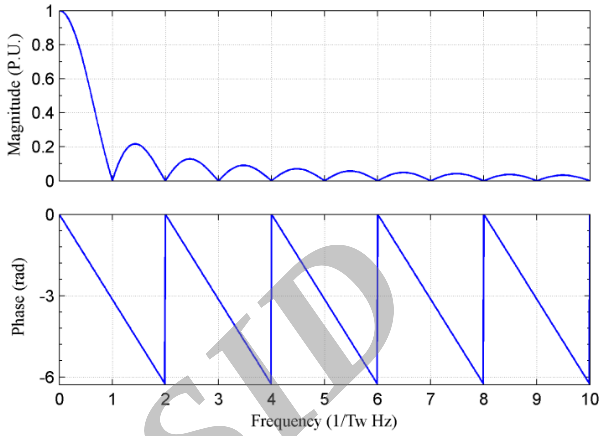

Fig. 2. Bode diagrames of MAF.

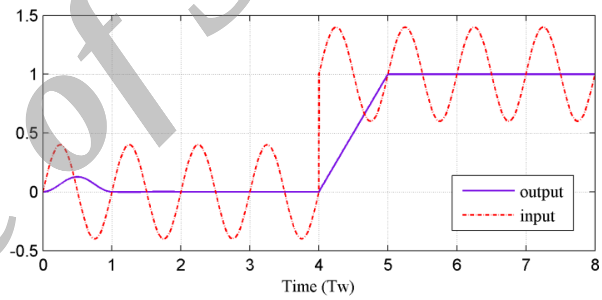

Fig. 3. Response of MAF for synthesized step and sinusoidal input signals.

A pure sinusoidal signal having period time *T* is filtered completely by MAF if $T_w = kT$, *k = 1, 2, 3*…. The MAF response for synthesized step and sinusoidal input signals are depicted in Fig. 3.

2. The Proposed method

The proposed method improves the basic structure of the demodulation-based PLLs, by using MAF as the LPF and establishing a feedback loop for frequency adaptation.

MAF can be tuned to eliminate oscillations precisely and fast in $T_w = T_{est}$ for input signals with the dc-offset or $T_w = T_{est}/2$ for offset-free inputs. MAF shows a better performance in the PLL structures than the LPN filter and other LPFs, because MAF removes the oscillations completely even if the cut-off frequency has a very small value. Consequently, the proposed approach can resolve the dc-offset problem.

Besides using the MAF, the proposed structure uses a feedback loop to estimate the frequency from the initial angle differences, as it is shown in Fig. 4. Therefore, the proposed structure contains an FLL

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and emulates the PLL function, leading to a more reliable performance.

If the initial frequency is not equal to the frequency of the input signal, the frequency difference (Δf) appears in θ_0' as

$$\begin{aligned} \theta_0' &= 2\pi(f - f_{\text{est}})t + (\theta_0 - \theta_{\text{tr},0}) \\ &= 2\pi\Delta f(t)t + (\theta_0 - \theta_{\text{tr},0}); \\ \text{s.t.} \quad \theta_{\text{tr},0} &= \int \Delta f(t)dt, \\ f_{\text{est}} &= f_0 + \Delta f(t). \end{aligned} \tag{10}$$

In this case, since the window length of MAFs do not exactly match to the exact frequency, MAF cannot completely eliminate all effects of the oscillations especially for the second-order component. Therefore, Sfα and Sfβ have high-order (particularly second-order) frequency components in this condition. This problem is resolved if the estimated frequency adapts to a precise value.

The proposed method uses the differentiation of θ_0' to estimate f from Δf. For this purpose, a sampling block with notch compensator (SBNC) is required to sample θ_0' having a saw-tooth nature (from –π to π) and high-order (second-order) oscillations. The sampling frequency of the feedback loop (fs2) should not be too high in order to avoid the propagation of oscillations in the loop, nor should be too low (at least 2fest) in order to have a fast enough response time. A reasonable sampling frequency, fs2, is recommended to be around 8fest to have a plausible estimation.

k is a proportional gain transforming differentiation of the SBNC output to an appropriate value for the frequency error (ef). If the sampling time of SBNC is Ts2, k is calculated as

$$k = 1/(2\pi T_{s2}). \tag{11}$$

The resulting frequency error is fed to a proportional integrative (PI) controller to generate an appropriate control signal to eliminate the error. The discrete transfer function of the PI controller is represented as

$$PI(z) = k_{\text{p}} + k_{\text{I}} T_{s,\text{PI}} \left(\frac{1}{z-1} \right). \tag{12}$$

The response time of the PI controller should be adapted with the fundamental frequency. The integral gain (k_{I}) is determined to be a function of f_{est}, the sampling frequency of PI controller ($T_{s,\text{PI}}$) is at least equal to T_{s2}, and the proportional gain (k_{p}) is a constant value.

Finally, a so-called "rate transition block" is used to interface the feedback to the main branch with a different sampling frequency. Fig. 5 shows the simplified model of the proposed method. This model is used for an estimation of the primary PI controller coefficients. Variable γ is equal to $(T_{\text{w}}/T_{s2})+1$ if the sampling time of the simplified model is

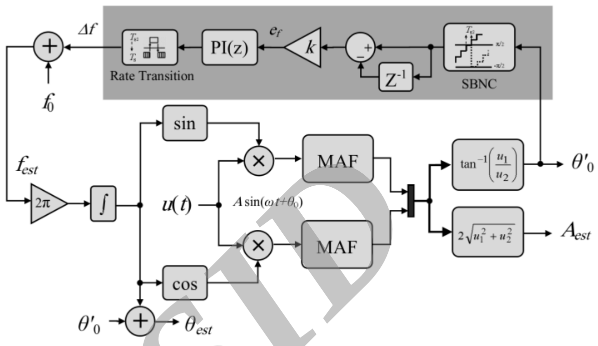

Fig. 4. Block diagram of the poroposed method.

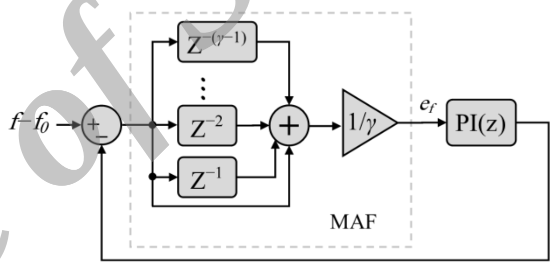

Fig. 5. Simplified model of the poroposed method.

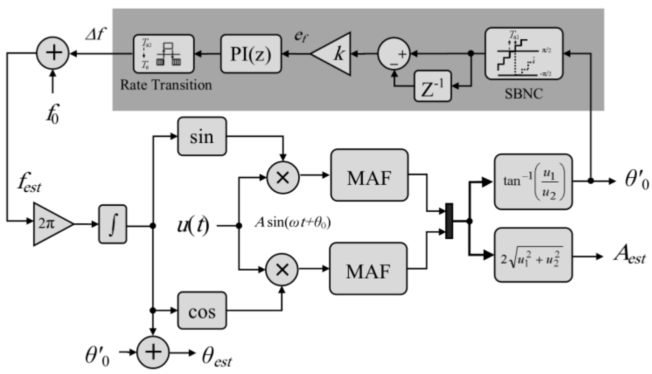

Fig. 6. Block diagram of the poroposed method.

T_{s2}. The primary coefficients are then tuned to provide an appropriate dynamic response for the original nonlinear system. In the present study, the controller coefficients are obtained to be as follows:

$$\begin{aligned} k_p &= 0.7/m, \\ \text{s.t.} \quad m &= f_{s2}/f_{\text{est}}, \end{aligned} \tag{13}$$

$$\begin{aligned} k_{\text{I}} &= 0.4n/T_{s2}, \\ \text{s.t.} \quad n &= T_{\text{est}}/T_{\text{w}}. \end{aligned} \tag{14}$$

3. Simulation

The proposed method was simulated for comparison with SOGI-FLL under distorted conditions including the harmonic distortion, dc-offset and step changes in amplitude, phase angle and frequency. For technical details on SOGI-FLL, readers are referred to [2].

A) Input signal with the dc-offset

To demonstrate the effectiveness of the proposed method against the dc component, three dc levels were considered as the value of the input dc-offset. Figs. 6, 7 and 8 show the simulation results having 1, 1.5 and 5 p.u. of dc-offset in the input signal.The proposed method has a superior performance than SOGI-FLL to cope the dc-offset. Fig. 6 suggests that the proposed method has a lower overshoot and shorter response time than SOGI-FLL. Moreover, Figs 7 and 8 show that, in contrast to SOGI-FLL, the proposed method provides a response with less oscillation and steady state error, as the value of dc-offset increases.

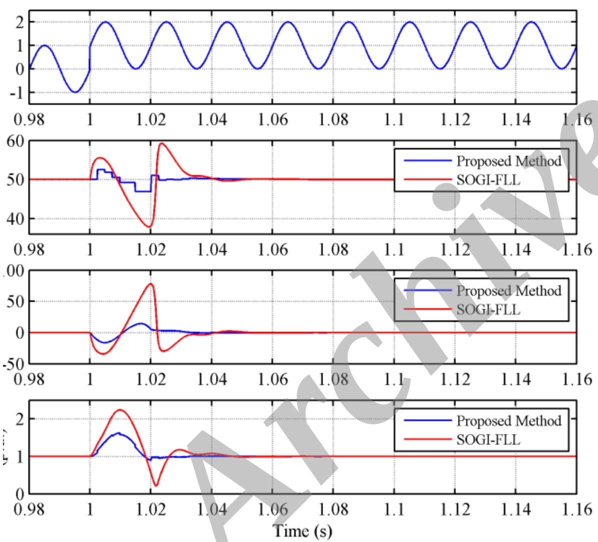

Fig. 7. Simulation results for the input signal with 1 p.u. of the dc-offset.

Input signal with the change in each fundamental parameter PLLs/FLLs that estimate all the fundamental parameters are adaptive against changes in the value of amplitude, phase angle and frequency.A step change is applied to the value of each fundamental parameter at time 1 Sec. Fig. 9 shows the dynamic responses of the proposed method compared with the SOGI-FLL performances for a –0.5 p.u. of amplitude change, –60˚ of phase jump, –5% of frequency change and –15% of frequency change in the input signal, respectively from left to right. The simulation was performed for both $T_W = T$ and $T_W = T/2$.

The results show that the proposed method outperforms the SOGI-FLL for the amplitude change. Both of the methods have a similar performance against the phase jump and the frequency changes with $T_W = T/2$; however, the settling time of the proposed method increases as the MAF window length increases.

B) Input signal with the harmonic distortion

In this stage, the input signal includes a harmonic distortion containing the 3th, 5th, 7th and 9th components, each having the amplitude of 0.05 p.u. The simulation results are shown in Fig. 10. The results for all fundamental parameters demonstrate that the proposed method has no steady state error in contrast with the SOGI-FLL that has considerable oscillations caused by the harmonic distortions.

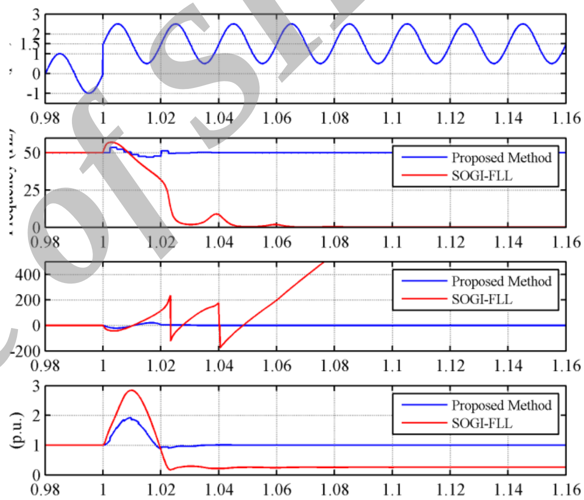

Fig. 8. Simulation results for the input signal with 1.5 p.u. of the dc-offset.

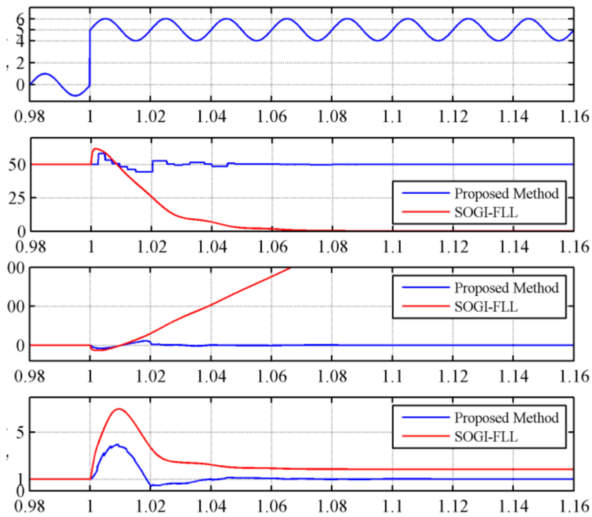

Fig. 9. Simulation results for the input signal with 5 p.u. of the dc-offset.

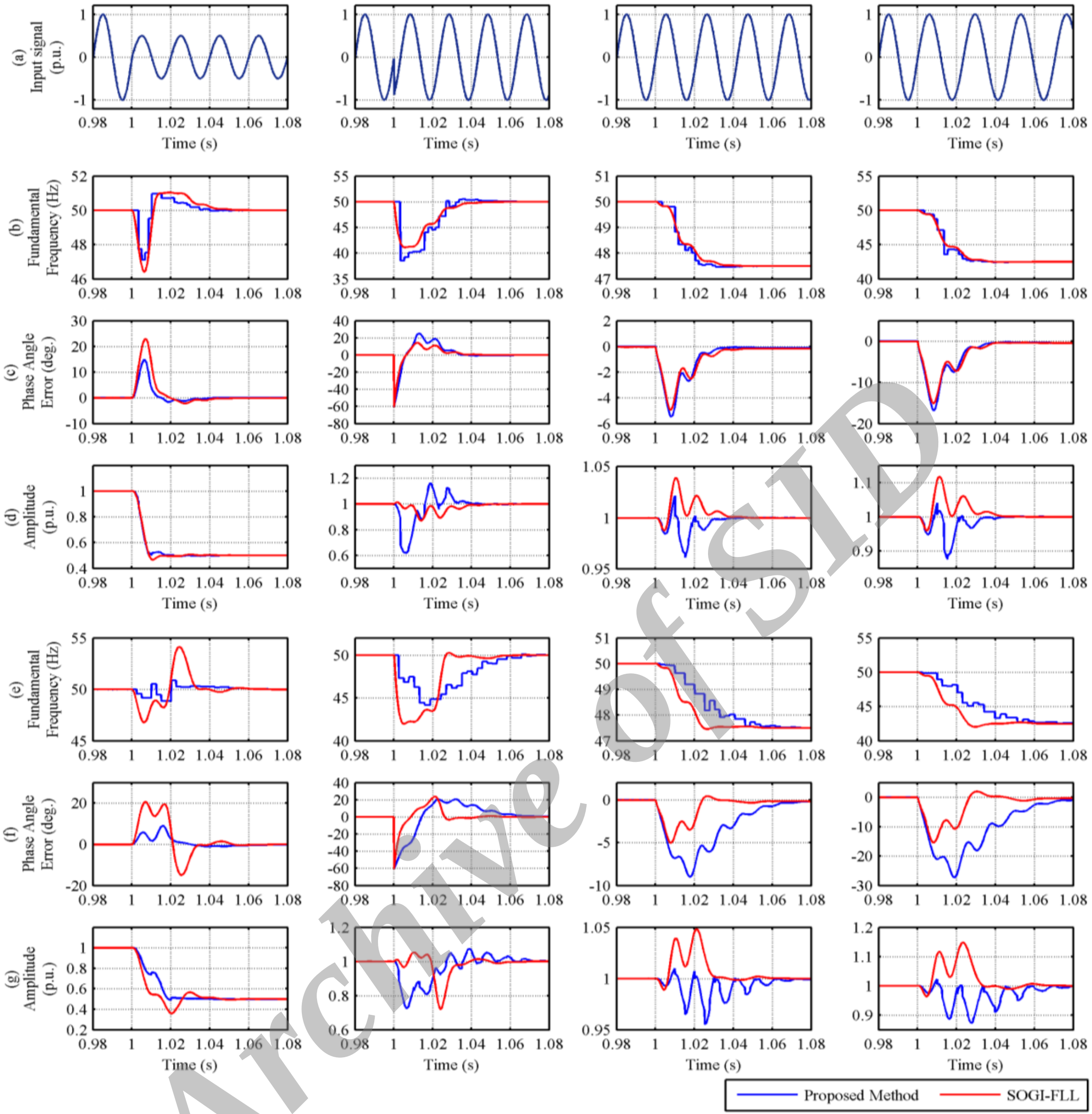

Fig. 10. Simulation results for the input signal with the change in each fundamental parameter; (a) input signal with a change in the amplitude (–0.5 p.u.), phase angle (–60˚) and frequrncy (–5% and –15%) respectively from the left column to right, (b, c, d) simulation results for $T_w = T/2$, (e, f, g) simulaiton results for $T_w = T$.

4. Conclusion

A novel method was proposed in this paper to estimate the fundamental parameters of distorted single phase signals. The estimator has two dynamics corresponding to the value of the MAF window length: one MAF setting is used when the input signal is free of the dc-offset and the even harmonic components that results in a fast and accurate estimation; and the other MAF setting is used against the dc-offset and all harmonic distortions, leading to a slower but more robust estimation.For input signals with no dc-offset and even harmonics, the proposed method offers a comparable performance to the SOGI-FLL, while in the presence of the dc-offset, harmonic distortion and amplitude changes, the proposed method outperforms the SOGI-FLL.The proposed structure includes the properties of PLL on phase tracking as well as FLL on frequency tracking eliminating the steady state error.

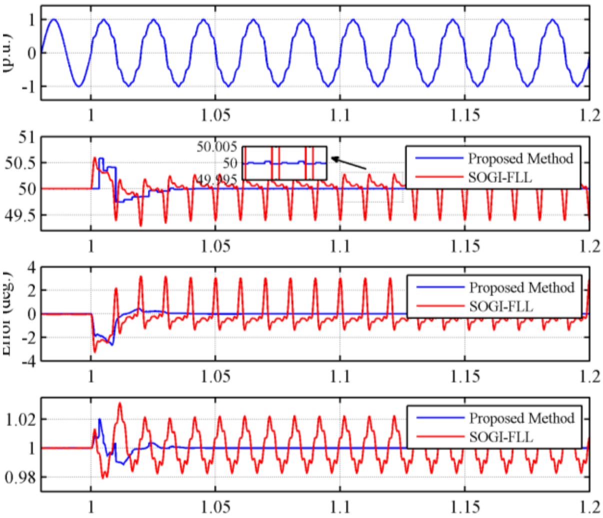

Fig. 11. Simulation results for the input signal with the harmonic distortion.

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