Robust Queen Bee Assisted Genetic Algorithm (QBGA) Optimized Fractional Order PID (FOPID) Controller for Not Necessarily Minimum Phase Power Converters

SURYA VARCHASVI DEVARAJ1, (Graduate Student Member, IEEE), MANAVAALAN GUNASEKARAN2, ELANGO SUNDARAM2, (Member, IEEE), MANIKANDAN VENUGOPAL2, SHARMEELA CHENNIAPPAN3, (Member, IEEE), DHAFER J. ALMAKHLES4, (Senior Member, IEEE), UMASHANKAR SUBRAMANIAM4, (Senior Member, IEEE), AND MAHAJAN SAGAR BHASKAR4, (Senior Member, IEEE)

1Department of Electrical Engineering, Indian Institute of Technology at Bombay, Bombay 400076, India
2Department of Electrical and Electronics Engineering, Faculty of Engineering, Coimbatore Institute of Technology, Coimbatore 641014, India
3Department of Electrical and Electronics Engineering, College of Engineering Guindy, Anna University, Chennai 600025, India
4Renewable Energy Laboratory, Department of Communications and Networks, College of Engineering, Prince Sultan University, Riyadh 11586, Saudi Arabia

Corresponding authors: Manavaalan Gunasekaran (manavaalan.g@cit.edu.in) and Dhafer J. Almakhles (dalmakhles@psu.edu.sa)

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ABSTRACT  Power electronic converters find application in diverse fields due to their high power conversion efficiency. Converters are often characterized by time response specifications, robustness and stability. Conventionally, converters employ the classic PID controller. The state space average linear time invariant model of a boost converter is known to be a non-minimum phase system. This paper demonstrates that the boost converter with a PID controller using the Queen Bee assisted Genetic Algorithm (QBGA) optimization is not robust to plant parameter variations. A fractional order PID controller based on QBGA optimization proposed here is shown to have improved robustness. The controller proposed here is applicable across converters, viz., buck, boost and buck-boost, equally.

INDEX TERMS  Boost converter, non-minimum phase system, QBGA, fractional order PID controller.

I. INTRODUCTION  
Power electronic converters are popular in use due to their high efficiency. Under plant parameter variations, closed loop controller design for converters with required regulation poses challenges. DC-to-DC converters often use buck converter, boost converter and buck-boost converter. In practice, design of controllers for converters are based on a simplified average linear time invariant (LTI) model even though the system is piece-wise linear [1]. In the process, the boost converter is represented by a state-space averaged model. Control of buck converters using PID controllers is satisfactory [2].

Control of boost converters using PID controller [3] performs well for input voltage and load changes; but not in the presence of plant parameter drifts. This is attributed to the non-minimum phase nature of the boost converter [4]. A few more converters which shows the non-minimum phase nature is provided in [5], that work presents the internal model control based PID tuning to maximize the bandwidth of converters.

General optimal tuning schemes for PID and fractional order PID (FOPID) controllers for unstable and integral plants are reported in [6]. Automatic tuning of optimum PID controllers based on a time-weighted integral performance criterion and integral of time error squared criterion are studied [7]. The Optimal Queen Bee assisted Genetic
Algorithm (QBGA) presented in [8] is used to tune the PID parameters in this paper.

The work of [9] presents the performance analysis of genetic algorithm (GA) and Queen Bee assisted GA tuned PI controller for shunt active power filter connected with complex loads. In which, QBGA tuned PI controller outperforms as compared to other tuning methods. In the present article, the considered plant is of non-minimum phase system and the performance of PID controller with respect to change in plant parameter is not satisfactory [4]. Which leads to explore another type of controller for not necessarily minimum phase power converters.

FOPID controller proposed in [10] provides flexibility and robustness to the system even in the time-delay systems [11] and under plant parameter variations. However the physical realization of FOPID is based on an integer order approximation of the controller transfer function [12], [13]. A procedure for conversion of FOPID to its integer order approximation using the state model is presented in [14]. This approximation is shown to result in a system of increased order, hence increasing the complexity in physical realization. A digital realization of fraction order controller for DC motor control is presented in [15]. There, the Digital FOPID controller for boost converter is used to obtain robustness.

The stability analysis of a fractional order system is based on Matignon’s stability theorem [16] and is quite different from that of an integer order system. This paper uses Matignon’s stability theorem for stability analysis of the boost converter using FOPID control.

Classical control of boost converter is presented in Section II. Section III comprises FOPID controller for boost converter, and a brief outline of QBGA as used for the optimal tuning of parameters. Simulations are presented and discussed in Section IV. Conclusions are drawn in Section V.

II. CLASSICAL CONTROL OF BOOST CONVERTER

A model for the boost converter and its control based on classical PID controller are discussed here.

The Circuit of boost converter is represented in Fig. 1. In which, QBGA tuned PI controller outperforms another type of controller for not necessarily minimum phase loads. In which, QBGA tuned PI controller outperforms another type of controller for not necessarily minimum phase loads.

A. THE BOOST CONVERTER MODEL

When S is ON, the governing equations of the boost converter are

\[
\begin{align*}
   v_i &= L \frac{di}{dt} + i_r r_l \\
   \frac{dv_c}{dt} &= \frac{1}{C} \left( \frac{v_c}{R_l + r_c} \right) \quad \text{and} \\
   v_o &= v_c + r_c \left( \frac{1}{R_l + r_c} \right) \\
   v_c &= -r_c i_c + R_l (i_l - i_c), \\
   i_c &= \frac{R_l i_l - v_c}{R_l + r_c} \\
   v_i &= L \frac{di_l}{dt} + i_r i_l + r_c \left( \frac{R_l i_l - v_c}{R_l + r_c} \right) + v_c \\
   \frac{dv_o}{dt} &= \frac{1}{C} \left( \frac{v_o}{R_l + r_c} \right) \quad \text{and} \\
   v_o &= r_c \left( \frac{R_l i_l - v_c}{R_l + r_c} \right) + v_c \\
   &= \frac{R_l i_l}{R_l + r_c} + \frac{R_l V_o}{R_l + r_c}.
\end{align*}
\]

Let \(x = [i_l \; v_c]^{T}\) be a state vector. Then, from (1) and (2), the piece-wise state-space On and OFF models for the circuit of Fig. 1 are

\[
\begin{align*}
   \dot{x} &= A_{ON} x + B_{ON} v_i \\
   v_o &= C_{ON} x \quad \text{for} \; T_{ON} \quad (3) \\
   \dot{x} &= A_{OFF} x + B_{OFF} v_i \\
   v_o &= C_{OFF} x \quad \text{for} \; T_{OFF}. \quad (4)
\end{align*}
\]

Here, \(A_{ON}, B_{ON}, C_{ON}, A_{OFF}, B_{OFF}\) and \(C_{OFF}\) are obtained from (1) and (2) and are given in Appendix.

Then, the average state-space model becomes

\[
\begin{align*}
   \dot{x} &= A x + B v_l \\
   v_o &= C x \quad (5)
\end{align*}
\]

with

\[
\begin{align*}
   A &= \frac{A_{ON} T_{ON} + A_{OFF} T_{OFF}}{T_{ON} + T_{OFF}}, \\
   B &= \frac{B_{ON} T_{ON} + B_{OFF} T_{OFF}}{T_{ON} + T_{OFF}}, \quad \text{and} \\
   C &= \frac{C_{ON} T_{ON} + C_{OFF} T_{OFF}}{T_{ON} + T_{OFF}}.
\end{align*}
\]

The average model of (5) is considered as the plant to design a controller. The simulations, however, are performed with the actual circuit of Fig. 1. The average model of (5) yields the transfer function

\[
\frac{V_o(s)}{D(s)} = \frac{V_o^2}{V_i} \left( 1 + \frac{s}{\omega_1} \right) \left( 1 - \frac{s}{\omega_2} \right) \left( 1 + \frac{s}{\omega_3} \right) \left( 1 - \frac{s}{\omega_4} \right)
\]

where \(v_o\) is the output voltage. With S OFF, the boost converter is described by the following:

\[
\begin{align*}
   v_i &= \frac{L}{R_c} \left( \frac{i_l}{R_l} - \frac{i_r}{r_c} \right) + r_c i_l + i_r \left( \frac{1}{R_l + r_c} \right) + v_c \\
   v_o &= r_c \left( \frac{1}{R_l + r_c} \right) + \frac{V_o}{r_c}.
\end{align*}
\]
where

\[
D = \frac{T_{ON}}{T_{ON} + T_{OFF}}
\]

\[
\omega_1 = \frac{1}{r C}
\]

\[
\omega_2 \approx \frac{R_L}{L} \left( \frac{\bar{V}_i}{\bar{V}_o} \right)
\]

\[
\omega_0 \approx \frac{1}{\sqrt{LC}} \frac{\bar{V}_i}{\bar{V}_o} \text{ and }
\]

\[
Q \approx \frac{r_L}{L + C(R_L + r_c)}
\]

\(\bar{V}_i\) and \(\bar{V}_o\) denote the nominal input and output voltages respectively.

In this paper the model parameters are chosen identical to those in [3] (vide Table 1) to facilitate a comparison of results obtained here with those in [3].

### TABLE 1. Nominal plant parameters.

| \(v_i\) (V) | \(P_i\) (kVz) | \(R_L\) (\(\Omega\)) | \(L\) (mH) | \(C\) (\(\mu F\)) | \(r_l\) (\(\Omega\)) | \(r_c\) |
|-------------|----------------|-----------------------|-----------|-------------------|-----------------|-------|
| 36          | 2              | 100                   | 33        | 1000              | 2               | 0.5   |

B. DRAWBACKS OF THE CLASSICAL PID CONTROLLER FOR BOOST CONVERTER

This section would like to examine the performance of PID controller for the boost converter deviates from the nominal parameter. The following section presents the results of PID controller proposed in [3] and the next section presents the results of PID controller whose gains are tuned based on Integral Square Error (ISE) optimization.

1) PID CONTROLLER TUNED BASED ON TRANSIENT RESPONSE OPTIMIZATION USING QBGA [3]

Let \(t_r\) : rise time, \(t_s\) : settling time, \(M_p\) : maximum peak over shoot and \(e_{ss}\) : steady state error. Also, let \(K_p\), \(K_i\) and \(K_d\) be the proportional, integral and differential gains of the PID controller respectively. The PID controller proposed in [3] solves the following constrained optimization problem:

Minimize \[ F = (1 + t_r)(1 + t_s)(1 + M_p)(1 + e_{ss}) \]

subject to \[ K_p \in [K_{p1}, K_{p2}], \]

\[ K_i \in [K_{i1}, K_{i2}], \] and

\[ K_d \in [K_{d1}, K_{d2}]. \]

In [3], the specifications for the nominal plant taken are \(t_r = 60 \text{ ms}, t_s = 1.3 \text{ s}, e_{ss} = 0.6\%, \text{ and } M_p = 0\%.

With a view to make this presentation self-contained, the implementation of QBGA [8] is shown in the flowchart of Fig. 7 in Appendix. Using QBGA, the optimal PID gains obtained are \(K_p = 0.326, K_i = 8.87\) and \(K_d = 0.012\).

Before presenting the simulation results, there is a need to pose and answer the following question: “Why seek robustness to parameter variation?” Significantly, the answer lies in the following as reported in [17]: (i) Temperature rise decreases the equivalent series resistance of an electrolytic capacitor \((r_c\) here); and (ii) Inductor coil resistance \((r_l\) here) increases with increase in temperature. The issue of the need for a robust controller thus settled. In this perspective, the simulation results are reported here.

With values as listed above (taken from [3]), the simulation results are captured in Figure 2. The following observations are of interest:

- The error in response is reasonably small with variations in \(L\). In fact, for \(t \in [2, 3] \text{ s}\), the maximum output disparity is as small as \(\pm 0.1 \text{ V}\).
- The disparity in response when \(r_l\) drifts in excess of 40\% of its nominal value is substantial. For \(t \in [2, 3] \text{ s}\), the corresponding maximum disparity in output is close to \(-6.3 \text{ V}\), which is quite large.
- The error oscillates and grows as \(r_c\) deviates by more than 40\% of its nominal value. For \(t \in [2, 3] \text{ s}\), it oscillates between \(\pm 2.8 \text{ V}\).

Clearly, robustness to parameter variations is critical to the controller performance, and that has not been provided by the scheme in [3].

FIGURE 2. Classical PID controller: (a) Plant output with nominal parameters. (b) Output disparity with change in \(L\). (c) Output disparity with change in \(r_l\). (d) Output disparity with change in \(r_c\).

2) QBGA BASED ISE OPTIMIZATION FOR PID CONTROLLER

This section would like to examine performance of PID controller whose gains are obtained by optimizing the integral square error. The gains of PID controller parameters are obtained by solving on the following optimization problem
using QBGA:

\[
\begin{align*}
\text{minimize} \quad & F = \int_0^\infty e^2(t)dt \\
\text{subject to} \quad & K_p \in [K_{p1}, K_{p2}], \\
& K_i \in [K_{i1}, K_{i2}], \\
& K_d \in [K_{d1}, K_{d2}].
\end{align*}
\]  

(7)

The parameters used to solve the above problem (7) are as follows: (i) Population size: 30; (ii) Bee structure: Binary; (iii) # Iterations: 50; (iv) Gains \( K_p \in (0, 5), K_i \in (0, 15) \) and \( K_d \in (0, 5) \).

The gains obtained from the above tuning are \( K_p = 4.07, K_i = 13.5, \) and \( K_d = 0.099 \). The corresponding simulation results are as in Figure 3. The following are the significant observations of the PID controller whose gains are obtained by solving the ISE optimization:

- The error in response is reasonably small with variations in \( L \). In fact, for \( t \in [2, 3] \) s, the maximum output disparity is as small as \( \pm 0.1 \) V.
- The disparity in response when \( r_t \) drifts in excess of 75% of its nominal value is substantial. For \( t \in [2, 3] \) s, the corresponding maximum disparity in output is close to \(-3.7 \) V, which is large but reduced as compared to result presented in Figure 2.
- The error oscillates and grows as \( r_c \) deviates by more than 40% of its nominal value. For \( t \in [2, 3] \) s, it oscillates between \( \pm 1.4 \) V, which is also reduced as compared to result presented in Figure 2.

Slight improvements are observed in the performance of PID controller tuned based on ISE as compared to PID controller tuned based on transient response [3]. However there exist a significant disparity in the output whenever the change in \( r_t \) and \( r_c \) are significant in boost converter.

### III. FOPID CONTROLLER FOR BOOST CONVERTER

To address the problem of lack of robustness of the controller as highlighted in Section II, this section presents the design of FOPID controller [12], [13] for the boost converter. FOPID controller output, \( U(S) \), is related to the controller input, \( E(s) \), as

\[
U(s) = \left( K_p + \frac{K_i}{s^\mu} + K_d s^\nu \right) E(s),
\]

where \( \lambda \) and \( \mu \) are positive real, thus admitting fractional order as different from the conventional PID control case. In this paper QBGA based optimization is used to tune these parameters, in addition to the PID controller gains.

#### A. QBGA BASED OPTIMIZATION FOR FOPID

The FOPID control parameters of (8) are obtained by solving the following optimization problem using QBGA:

\[
\begin{align*}
\text{minimize} \quad & F = \int_0^\infty e^2(t)dt \\
\text{subject to} \quad & K_p \in [K_{p1}, K_{p2}], \\
& K_i \in [K_{i1}, K_{i2}], \\
& K_d \in [K_{d1}, K_{d2}], \\
& \lambda, \mu \in (0, 1].
\end{align*}
\]  

(9)

The QBGA-based optimal tuning of the FOPID parameters to be used in a boost converter is as in the schematic of Figure 4, which is self-explanatory.

![Figure 4](image-url)  

**FIGURE 4.** Schematic of QBGA-based optimal tuning of FOPID parameters for boost converter.

### IV. SIMULATIONS AND DISCUSSIONS

For implementing the QBGA-based optimization for the problem (9), the parameters used in this paper are as follows: (i) Population size: 30; (ii) Bee structure: Binary; (iii) # Iterations: 100; (iv) \( K_p, K_i, K_d \in (0, 10] \); and \( \lambda, \mu \in (0, 1] \).

The tuned output values from the QBGA optimization are \( K_p = 3.5, K_i = 6, K_d = 0.001, \lambda = 0.8, \mu = 0.3 \). The corresponding simulation results are as in Figure 5. The following are the significant observations and comparisons with the results using the classical PID controller:

- The error in response with drift in \( L \) not only remains small but has reduced even further; it has been contained between \( \pm 0.1 \) V for \( t \in [2, 3] \) s.
- The disparity in response when \( r_t \) drifts in excess of 40% of its nominal value has greatly reduced. For \( t \in [2, 3] \) s, the maximum disparity in output is close to \(-0.96 \) V.
- Though the error oscillates as \( r_c \) deviates by more than 40% of its nominal value, the oscillations for \( t \in [2, 3] \) s
are only in the positive side; more importantly, the magnitude is less than 0.27 V.

- The steady state error starts increasing whenever the load resistance falls below 70Ω and the input voltage falls below 30 V. Which is also similar to that of boost converter with PID controller.

A comparison of the maximum absolute disparities in the responses with the classical PID controller and the QBGA based FOPID controller proposed here for the interval \( t \in [2, 3] \) s is provided in Figure 6. The percentage of output disparity with respect to desired output at steady state \((t \in [2, 3])\) s for both PID and FOPID controllers are provided in Table 2. The enhancement in robustness with the present QBGA-based FOPID controller as claimed clearly stands out.

A. NOTE ON PRACTICAL IMPLEMENTATION OF FOPID CONTROLLER

Though the FOPID controller design for the boost controller significantly improves robustness to plant parameter variations in principle, its practical implementation is saddled with having to ‘realise’ the fractional order system with an approximate transfer function of high order. This is necessary to match the frequency response in the band of significance by a recursive procedure [12], [13]. In fact, the FOPID controller obtained here corresponds to a transfer function of order 23. Analog implementation of such high order systems is known to be fraught with tolerance effects of individual components. Hence implementation of discrete version of integer order approximated transfer function is necessary. That provides the direction for future work.

V. CONCLUSION

The performance of the classical PID controller for the boost converter falls short in terms of robustness to parameter variations, which are a reality. The non-minimum phase nature of the boost converter causes this infirmity. There is a need for a controller which is agnostic to the plant being minimum phase or non-minimum phase. The fractional order PID controller proposed here is based on optimal tuning of the parameters using a queen bee genetic algorithm optimization. The performance of the classical PID controller for a boost converter is compared with that of the one proposed here.
The robustness of the boost converter with this controller is demonstrated through simulations. Though this paper specifically discusses the design of a robust controller for a boost converter, the controller proposed here works equally well for buck and buck-boost converters too. Practical implementation of a FOPID controller requires a digital implementation; that suggests the vistas for the future.

APPENDIX

\[ A_{ON} = \begin{bmatrix} \frac{-r_I}{L} & 0 & 0 \\ 0 & \frac{-1}{C(R_L + r_c)} \end{bmatrix} \]

\[ A_{OFF} = \begin{bmatrix} \frac{1}{L} & 0 & 0 \\ 0 & \frac{-L}{C(R_L + r_c)} \end{bmatrix} \]

\[ B_{ON} = B_{OFF} = \begin{bmatrix} 1 \\ L \\ 0 \end{bmatrix} \]

\[ C_{ON} = \begin{bmatrix} 0 \\ \frac{R_L}{R_L + r_c} \end{bmatrix} \]

\[ C_{OFF} = \begin{bmatrix} \frac{-R_I}{L} \\ \frac{-R_I}{L} + \frac{R_L}{R_L + r_c} \end{bmatrix} \]

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SURYA VARCHAVSI DEVARAJ (Graduate Student Member, IEEE) received the master’s degree from the Coimbatore Institute of Technology, in 2015, with a specialization in applied electronics. He is currently pursuing the Ph.D. degree with the Indian Institute of Technology Bombay (IIT Bombay). He worked as an Assistant Professor with the CMR Institute of Technology, Bengaluru, from 2015 to 2018. His research interests include control systems, converters, integrated circuit design, and front-end circuits for sensors.

ELANGO SUNDARAM (Member, IEEE) was born in Tamil Nadu, India, in June 1972. He received the B.E. degree in electrical and electronics engineering and the M.E. degree in applied electronics from the Coimbatore Institute of Technology, Coimbatore, and the Ph.D. degree in electrical engineering from Anna University, Chennai, in 2016. He is currently an Associate Professor with the Department of Electrical and Electronics Engineering, Coimbatore Institute of Technology. He has published more than ten international journals and 25 international conferences. His current research interests include power electronics applications into power system and smart grid, power quality, and soft computing techniques.

MANAVAALAN GUNASEKARAN received the bachelor’s degree in electrical engineering from the University of Madras, Chennai, India, in 2002, the master’s degree in applied electronics from the Coimbatore Institute of Technology, Coimbatore, India, in 2005, and the Ph.D. degree in electrical engineering from the Indian Institute of Technology at Kanpur, Kanpur, India, in 2015. In 2017, he joined the Coimbatore Institute of Technology, India, as a Faculty Member. His research interests include applications of control systems, smart electrical vehicles, robotics, and automation and soft computing techniques.
Recent research outcomes related to renewable energy systems, electric vehicles, and power system automation have been presented in leading international conferences in India and abroad. He has expertise in energy auditing and energy optimization. His research interests include modeling and synthesis of efficient lossless power electronic conversion circuits to cater to power interfaces between PV/Wind sources and grid, development of soft-computing-based control strategies for smart power systems, and embedded electronics.

**MANIKANDAN VENUGOPAL** is currently serving as a Professor for the Department of Electrical and Electronics Engineering (EEE), Coimbatore Institute of Technology, Coimbatore, India, for more than two decades. He has guided nine Ph.D. scholars, in the last ten years, from various Indian funding agencies, like AICTE, DST, and MHRD for executing various research projects, conduction of faculty development programs and infrastructure developmental activities in the institute campus and in the neighboring ecosystems. His research interests include modeling and synthesis of efficient lossless power electronic conversion circuits to cater to power interfaces between PV/Wind sources and grid, development of soft-computing-based control strategies for smart power systems, and embedded electronics.

**SHARMEELA CHENNAIAPPAN** (Member, IEEE) received the B.E. degree in electrical and electronics engineering, the M.E. degree in power systems engineering from Annamalai University, Chidambaram, and the Ph.D. degree in electrical engineering from the College of Engineering, Guindy (CEG), Anna University, Chennai. She has a teaching/research and consultancy experience of 19 years in the areas of power quality and power systems. She was the Assistant Director, Centre for Entrepreneurship Development, CEG, Anna University, from 2015 to 2018. She has done a number of consultancies on renewable energy systems, such as solar photovoltaic (SPV) power systems, power quality measurements, and design of compensators for industries. She currently holds the positions of an Associate Professor and a Professor In-Charge in power engineering and management with the Department of Electrical and Electronics Engineering, CEG, Anna University, Chennai. She has coordinated and organized several short-term courses on power quality for Tamil Nadu State Electricity Board Engineers, TN, India. She has also delivered several invited talks and trained more than 1000 engineers on the importance of power quality, power quality standards, and design of SPV power systems, for more than 12 years in leading organizations, such as CII, FICCI, CPRI, MSME, GE (Alstom), and APQI. She has received the grant from CTDT, Anna University, for a two-year project on “Energy Efficient Solar-Based Lighting System for Domestic Application,” in 2011. She has also received a research grant from AICTE–RPS, New Delhi, India, on “Smart EV Charging Station,” in 2020. She has authored over 30 journal articles in refereed international journals and more than 60 papers in international and national conferences. She has also authored/coauthored/editing five book chapters, edited two books, and authored one book. Her research interests include power quality, power electronics applications to power systems, smart grid, energy storage systems, renewable energy systems, electric vehicle, battery management systems, and electric vehicle supply equipment. She is a fellow of the Institution of Engineers, India, a Life Member of ISTE, the Central Board of Irrigation and Power (CBIP), New Delhi, India, and SSI, India.

**UMASHANKAR SUBRAMANIAM** (Senior Member, IEEE) has more than 16 years of teaching, research, and industrial research and development experience. He worked as an Associate Professor and the Head of the VIT, Vellore, and a senior research and development engineer and a senior application engineer in the field of power electronics, renewable energy, and electrical drives. He is currently an Associate Professor with the Renewable Energy Laboratory, College of Engineering, Prince Sultan University, Riyadh, Saudi Arabia. Under his guidance, 24 P.G. students and more than 25 U.G. students completed the senior design project work. Six Ph.D. scholars also completed Ph.D. theses as a Research Associate. He is also involved in collaborative research projects with various international and national level organizations and research institutions. He has published more than 250 research articles in national and international journals and conferences. He has also authored/coauthored/contributed 12 books/chapters and 12 technical articles on power electronics applications in renewable energy and allied areas. He is a Senior Member of PES, IAS, PSES, YP, and ISTE. He was an Executive Member, from 2014 to 2016. He received the Danfoss Innovator Award-Mentor, from 2014 to 2015 and from 2017 to 2018, and the Research Award from the VIT University, from 2013 to 2018. He also received the INAE Summer Research Fellowship, for the year 2014. He was the Vice Chair of IEEE MAS Young Professional by the IEEE Madras Section, from 2017 to 2019. He has taken charge as the Vice Chair of the IEEE Madras Section and the Chair of the IEEE Student Activities, from 2018 to 2020. He is an Editor of *Heliyon* (Elsevier) and various other reputed journals. He is an Associate Editor of *IEEE Access*.

**MAHAJAN SAGAR BHASKAR** (Senior Member, IEEE) received the bachelor’s degree in electronics and telecommunication engineering from the University of Mumbai, Mumbai, India, in 2011, the master’s degree in power electronics and drives from the Vellore Institute of Technology (VIT University), India, in 2014, and the Ph.D. degree in electrical and electronic engineering, University of Johannesburg, South Africa, in 2019. He is currently with the Renewable Energy Laboratory, Department of Communications and Networks Engineering, College of Engineering, Prince Sultan University, Riyadh, Saudi Arabia. He has authored more than 100 scientific articles, with particular reference to X. Y. converter family, multilevel DC/DC and DC/AC converter, and high gain converter.

**DHAFER J. ALMAKHLES** (Senior Member, IEEE) received the B.E. degree in electrical engineering from the King Fahd University of Petroleum and Minerals, Dhahran, Saudi Arabia, in 2006, and the master’s degree (Hons.) in electrical engineering and the Ph.D. degree in electrical and computer engineering from The University of Auckland, Auckland, New Zealand, in 2011 and 2016, respectively. Since 2016, he has been with Prince Sultan University, Riyadh, Saudi Arabia, where he is currently the Chairman of the Department of Communications and Networks Engineering, and the Director of the Science and Technology Unit and the Intellectual Property Office. He is also the Leader of the Renewable Energy Research Team and Laboratory.