IMPROVEMENT OF POWER CONVERSION EFFICIENCY ISOLATED BIDIRECTIONAL AC-DC CONVERTER FOR A DC DISTRIBUTION SYSTEM USING FUZZY LOGIC CONTROLLER

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Abstract-
A high-efficiency isolated bidirectional ac–dc converter is proposed for a 380-V dc power distribution system to control bidirectional power flows and to improve its power conversion efficiency using fuzzy logic controller. To reduce the switches’ losses of the proposed nonisolated full-bridge ac–dc rectifier using an unipolar switching method, switching devices employ insulated-gate bipolar transistors, MOSFETs, and silicon carbide diodes. Using the analysis of the rectifier’s operating modes, each switching device can be selected by considering switch stresses. A simple and intuitive frequency detection method for a single-phase synchronous reference frame-phase-locked loop (SRF-PLL) is also proposed using a filter compensator, a fast period detector, and a finite impulse response filter to improve the robustness and accuracy of PLL performance under fundamental frequency variations. In addition, design and control methodology of the bidirectional full-bridge LLC resonant converter is suggested for the galvanic isolation of the dc distribution system. A dead-band control algorithm for the bidirectional ac–dc converter is developed to smoothly change power conversion directions only using output voltage information. Simulation results will verify the performance of the proposed methods using fuzzy logic controller.

Index Terms—AC–DC boost rectifier, bidirectional isolated converter, LLC resonant converter dc distribution system, fuzzy logic controller.

INTRODUCTION
DC Distribution system is one of important future power systems to save energy and to reduce CO2 emission because it can improve the efficiency of systems due to the reduction of the number of power conversion stages [1]–[3]. Especially, the dc distribution system for a residential house using dc home appliances can allow the flexibility of merging many renewable energy sources because most of the output of renewable energy sources is dc. The overall system configuration of the proposed 380-V dc distribution system is shown in Fig. 1. In order to balance the power flow and to regulate the dc-bus voltage, the dc distribution system requires an isolated bidirectional ac–dc converter to interface between dc bus and ac grid. It usually consists of a non isolated bidirectional ac–dc rectifier [4]–[9] for grid-connected operation and an isolated bidirectional dc–dc converter to interface dc bus and dc link of the rectifier [10]–[18]. The single-phase nonisolated bidirectional rectifier typically consists of a conventional full-bridge structure. It has two sinusoidal pulsewidth modulation (SPWM) methods such as the bipolar and...
the unipolar switching modes. One of the disadvantages of the bipolar switching mode is the need of a large inductor to reduce the input current ripple because the peako-peak voltage of the inductor is more than twice the unipolar switching mode. If the full-bridge rectifier operates in the unipolar switching mode, inductance for a continuous current mode (CCM) power factor correction (PFC) operation can be reduced. One of full-bridge rectifier legs in the unipolar switching mode is operated at a line frequency while the other one is modulated at a switching frequency. However, the unipolar switching mode rectifier using conventional switching devices including a normal antiparallel diode causes high reverse recovery current and turn-on switching noise. The switching and the conduction losses in the bidirectional rectifier are the main cause of decreasing power conversion efficiency. The phase estimation, so-called phase-locked loop (PLL), is required to control the bidirectional ac–dc rectifier; especially, the phase information of supply voltage is mandatory to generate a current reference. One of the popular PLL methods is synchronous reference frame (SRF-PLL) which uses a rotating reference frame for tracking a phase angle. However, the conventional SRF-PLL has a weak point of frequency tracking performance because it uses the constant angular frequency of a fundamental component. It can cause a tracking error in the PLL operation when the fundamental frequency changes to the different value of the constant angular frequency. Numerous methods for improved PLL have been presented and introduced in the literature [19]–[23]. Even though they have good performance against the frequency distortion, their algorithms are complicated to be implemented for various applications. Another PLL method has been proposed using a simple frequency detector and trigonometric calculations without any linear phase detector; however, this method requires the halfcycle data acquisition of the fundamental frequency to detect the exact line frequency [24]. Therefore, a more simple, faster, and more intuitive frequency detection method should be upgraded for improving the performance of the single-phase bidirectional rectifier. Some isolated full-bridge bidirectional dc–dc converter topologies have been presented in recent years. A boost fullbridge zero-voltage switching (ZVS) PWM dc–dc converter was developed for bidirectional high-power applications. However, it needs extra snubber circuits to suppress the voltage stress of the switches [25]–[28]. A bidirectional phase-shift full-bridge converter was proposed with high-frequency galvanic isolation for energy storage systems [29], [30]. This converter can improve power conversion efficiency using a zero-voltage transition feature; however, it requires input voltage variations to regulate constant output voltage because this topology can only achieve the step-down operation. A bidirectional full-bridge LLC resonant converter was introduced for a UPS system without any snubber circuits [17]. This topology can operate under soft-switching conditions of primary switches and secondary rectifier. In addition, the topology confines voltage stresses without any clamp circuits. However, application of this converter showed different operations between transformer’s turn ratio and the difference of resonant networks. Therefore, the novel design guides suitable for a 380-V dc distribution system should be proposed. In this paper, the high-efficiency isolated bidirectional ac–dc converter system with several improved techniques will be discussed to improve the performance of a 380-V dc distribution system. In order to increase the efficiency of the nonisolated full-bridge ac–dc rectifier, the switching devices are designed by using insulated-gate bipolar transistors (IGBTs) without an antiparallel diode, MOSFETs, and silicon carbide (SiC) diodes. Through the analysis of operational modes, each switch is selected by considering switch stresses. The major novelty of the proposed PLL is the suggestion of a simple and intuitive frequency detection method for the single-phase SRF-PLL using an advanced filter compensator, a fast quad-cycle detector, and a finite impulse response (FIR) filter. Finally, design guides and gain characteristics of the bidirectional full-bridge LLC resonant converter with the symmetric structure of the primary inverting stage and secondary rectifying stage will be discussed for a 380-V dc distribution system. Experimental results will verify the performance of the proposed methods using a 5-kW prototype converter.
II. CIRCUIT CONFIGURATION OF THE PROPOSED ISOLATED BIDIRECTIONAL AC-DC CONVERTER

Fig. 2 shows the circuit configuration of the proposed isolated bidirectional ac–dc converter. It consists of the singlephase bidirectional rectifier for grid interface and the isolated bidirectional full-bridge CLLC resonant converter for galvanic isolation. To control the proposed converter, a single digital signal processor (DSP) controller (TMS320F28335) was used. The power flow directions in the converter are defined as follows: rectification mode (forward direction of power flow) and generation mode (backward direction of power flow).

The switching method of the proposed single-phase bidirectional rectifier is unipolar SPWM. In order to reduce the switching losses caused by the reverse recovery current in therectification mode, the high-side switches of the proposed rectifier are composed of two IGBTs without antiparallel diodes (S1 and S3) and two SiC diodes (DS1 and DS3). The lowside switches are composed of two MOSFETs (S2 and S4) for reducing conduction loss and for using ZVS operation in the generation mode. The detailed circuit operation of the proposed bidirectional rectifier and advanced PLL method will be discussed in Section III.

The proposed bidirectional full-bridge CLLC resonant converter has the full-bridge symmetric structure of the primary inverting stage and secondary rectifying stage with a symmetric transformer. Using the high-frequency transformer, the converter can achieve galvanic isolation between the primary side and the secondary side. The transformer Tris modeled with the magnetizing inductance Lm and the transformer’s turn ratio of 1:1. The leakage inductance of the transformer’s primary and secondary windings is merged to the resonant inductor Lr1 and Lr2, respectively. The resonant capacitors Cr1 and Cr2 make automatic flux balancing and high resonant frequency with Lr1 and Lr2. The detailed analysis and design guides of the proposed converter will be discussed in Section IV.

III. NONISOLATED AC–DC BIDIRECTIONAL RECTIFIER

High-power rectifiers do not have a wide choice of switching devices because there are not many kinds of the switching devices for high-power capacity. Generally, the full-bridge rectifier in high-power applications consists of the same four devices: IGBT modules or intelligent power modules (IPMs) are chiefly used. This modular device have antiparallel diodes, which have fast recovery characteristics. A fast recovery diode (FRD) has a small reverse recovery time trr. When the fullbridge rectifier operates, the time trrcauses a reverse recovery current which increases power loss and EMC problems. Therefore, soft-switching techniques using additional passive or active snubber circuits have been proposed [31]–[33]. Even though these methods require a relatively large number of passive or active components, which decrease the reliability of the rectifier system and increase system cost, the soft-switching techniques in the high-power rectifier are a unique solution for reducing the reverse recovery problems. On the other hand, a medium-power rectifier system around 5 kW for a residential house or building has a wide selection of switching devices such as discrete-type IGBTs and MOSFETs. Especially, commercial IGBTs without the antiparallel diode can be selected to replace antiparallel FRDs to SiC diodes. In theory, the SiC diode does not have trr.

Therefore, the combination of IGBTs without the antiparalleldiode and the SiC diodes is another viable solution to reduce the reverse recovery problems. In the bidirectional rectifier using the unipolar switching method, the turn-on period of low-side switches increases one and half times more than the turn-on period of the high-side switches. Therefore, the low-side switches should be chosen to consider low conduction losses for increasing the power conversion efficiency. The latest generation of MOSFETs employing superjunction technology can achieve extremely low on-resistance RDS.on. It is better than the latest very low VC E.

Fig. 3. Operating modes of the proposed bidirectional ac–dc rectifier in the rectification mode: (a) Mode 1, (b) Mode 2, (c) Mode 3, (d) Mode 4, and (e) Mode 5.

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IGBTs in the viewpoint of the conduction loss when the same current flows into the switching devices. Therefore, MOSFETs are suitable for the low-side switches in the unipolar switching method.

A. Consideration for Reverse Recovery Losses in a Rectification Mode

In the rectification mode, the bidirectional rectifier has five operating modes in a single switching cycle. The circuit operations in the positive half period of the input voltage are shown in Fig. 3. The dark lines denote conducting paths for each state. The theoretical waveforms of the proposed rectifier are given in Fig. 4.

At time $t_0$, the low-side switch $S_2$ turns ON. At this time, if $DS_1$ is FRD, $DS_1$ cannot immediately turn OFF because of its reverse recovery process. This simultaneous high reverse recovery current causes an additional switching loss on $S_2$. The reverse recovery current increases the current stress on the low-side switches and decreases the EMI performance of the rectifier.

To solve this reverse recovery problem, the high-side switches of the proposed circuit should use IGBTs without antiparallel diodes and SiC diodes as antiparallel diodes of the IGBTs. Even though the reverse recovery current is not completely zero in a practical manner, it is significantly reduced as compared with the FRD operation. After the discharge operation of the dc-link’s energy, the antiparallel diode including the low-side switch $S_2$ will be conducted by freewheeling operation using inductor’s energy as shown in Mode 2. During this period, the energy stored in the output capacitance of $S_2$ can be fully discharged. In Mode 3, $S_2$ turns ON under the ZVS condition. Through these operation modes, the turn-on losses in the low-side switches can be reduced.

When the high-side switch $S_1$ turns ON in Mode 5, the antiparallel diode of $S_2$ cannot immediately turn OFF because of poor reverse recovery performance of the MOSFET’s antiparallel diode. It causes an additional switching loss on $S_1$ through the reverse recovery current.

B. Consideration for Switching Losses in a Generation Mode

In the generation mode using the same switching pattern as the rectification mode, the proposed bidirectional rectifier has five operating modes in a single switching cycle. The circuit operations in the positive half period of the input voltage are shown in Fig. 5. After the discharge operation of the dc-link’s energy, the antiparallel diode including the low-side switch $S_2$ will be conducted by freewheeling operation using inductor’s energy as shown in Mode 2. During this period, the energy stored in the output capacitance of $S_2$ can be fully discharged. In Mode 3, $S_2$ turns ON under the ZVS condition. Through these operation modes, the turn-on losses in the low-side switches can be reduced.

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rectification mode: (a) Mode 1, (b) Mode 2, (c) Mode 3, (d) Mode 4, and (e) Mode 5

switching and conduction losses. However, if IGBTs are used for the low-side switches, the ZVS operation is not significant to reduce their switching loss. There are also reverse recovery losses through the reverse recovery characteristics of the antiparallel diode. They can increase the turn-off switching losses through the IGBT’s tailing current. If the reverse recovery loss through the MOSFETs is significant, it can be overcome employing the inverted switching pattern in the generation mode. The operation mode of the inverted switching pattern is shown in Fig. 6. In this operation, the switching pattern is perfectly inverted and the turn-on period of the high-side switches is one and half times longer than the turn-on period of the low-side switches. In Mode 5, the low-side switch $S_4$ turns ON. At the same time, the reverse recovery current should be limited by the SiC diode $DS_3$. Using this switching pattern in the generation mode, there are no benefits of the ZVS operation. However, the reverse recovery losses can be significantly reduced as compared with the same switching pattern does not affect the power conversion and control performance of the ac–dc rectifier.

C. Consideration for Conduction Losses

In the unipolar switching method, the turn-on period of low-side switches is one and half times longer than the turn-on period of high-side switches. To analyze the conduction loss of MOSFETs as the low-side switches, the circuit operation of the proposed bidirectional rectifier assumes that inductor $L_1$ is sufficiently large to operate CCM and the ac input current is a perfectly sinusoidal waveform. Generally, the conduction loss in MOSFETs can be calculated using the rms current passing through MOSFET's $R_{DS.on}$. Since the turn-on period of the low-side switches is 75% of the fundamental period of the ac input current, the rms current of the low-side switches can be calculated using the following equation:

$$I_{rms,low} = \sqrt{\frac{1}{2\pi} \int_0^{\frac{\pi}{2}} (I_{in,p} \sin \omega t)^2 dt}.$$  \hspace{1cm} (1)

Through the aforementioned assumptions, the peak current $I_{in,p}$ and the average current $I_{in,av}$ of the ac input can be derived as follows:

$$I_{rms,low} = \sqrt{\frac{1}{2\pi} \int_0^{\frac{\pi}{2}} (I_{in,p} \sin \omega t) dt} \hspace{1cm} (1)$$

Through the aforementioned assumptions, the peak current $I_{in,p}$ and the average current $I_{in,av}$ of the ac input can be derived as follows:

$$I_{in,p} = \sqrt{\frac{P_{out}}{\eta V_{in,av}}} \hspace{1cm} (2)$$

$$I_{in,av} = \frac{2}{\pi} I_{in,p} \hspace{1cm} (3)$$

where $\eta$ is the power conversion efficiency. Using the aforementioned equations, the total conduction losses of the rectifier's switches in the rectification can be calculated as follows:

$$P_{con-rec} = 2(I_{rms,low})^2 R_{DS.on} + 2\left(\frac{1}{2}I_{in,av}\right) V_F = \frac{3}{4} I_{in,p} R_{DS.on} + \frac{1}{\pi} I_{in,p} V_F \hspace{1cm} (4)$$

where the conduction loss in the rectification mode is $P_{con-rec}$ and $V_F$ is the forward voltage drop of the
high-side SiC diode, respectively. In comparison to using IGBTs for the low-side switches, Fig. 7 shows the calculated conduction losses using (4). The low-side switches use MOSFETs, which have extremely low $R_{DS(on)}$ (IXKR47N60C5) and the latest IGBTs (IKW75N60T), which have the very low collector–emitter threshold voltage $V_C E$. The high-side switches use SiC diodes (C3D20060D) as the antiparallel diode and IGBTs without antiparallel diode (IGW75N60T). The typical values of $R_{DS(on)}, V_F$, and $V_C E$ are obtained from their datasheets. Since $V_F$ and $V_C E$ are almost the same value, the conduction loss in the generation mode is expected to be nearly the same as the conduction loss in the rectification mod, $P_{con-rec}$. In Fig. 7, the conduction losses of MOSFETs are less than the conduction losses of IGBTs under the load condition of 6.5 kW or less. Under the light-load condition, MOSFETs is better than IGBTs in the view point of the power conversion efficiency. At the rated load (5 kW), the difference of the conduction losses between MOSFETs and IGBTs is about 12 W. If two MOSFETs are used in parallel, the conduction losses in the rated load will be reduced about 30 W.

### D. SRF-PLL With Enhanced Frequency Estimator

For the application of the grid-interactive power converters, the phase estimation of the supply voltage is required to control the entire power system; especially, the phase information of the supply voltage is mandatory to generate the current reference. The phase estimation method requires robustness to noise and disturbance from grid, accuracy to the fundamental frequency variation and harmonic distortion, and easy implementation using analog or digital platforms. The SRF-PLL is one of the popular methods. It has a weak point of a frequency tracking performance. In Fig. 8, the constant angular frequency of the fundamental component $\omega_s$ is used as a feedforward term for compensating the phase angle tracking. It can cause a tracking error in the PLL operation when the fundamental frequency changes to the different value of the constant $\omega_s$. To track unexpected fundamental frequencies, the SRF-PLL requires a frequency estimator instead of the constant angular frequency.

The SRF-PLL method uses the $\pi/2$ phase-shifted wave of the supply voltage using the all-pass filter (APF). Therefore, the PLL has four zero crossing points in a period of the supply voltage and they are located at every $\pi/2$. It means that the frequency information can be obtained at every quarter of the period. However, the conventional research has updated the frequency information at every half of the period. Consequently, the proposed frequency detection algorithm illustrated in Fig. 9(a) can update the variation of the supply frequency two times faster than the conventional one. In the algorithm, there are two time zones to calculate the supply period: $T_c$ and $T_d$ where $T_c$ is the time duration between two zero crossing points and $T_d$ is the time duration near the zero crossing point. $T_d$ is introduced to implement a practical PLL system using the proposed algorithm since noises in the power stage and sensors make the exact detection of the zero crossing points imperfect. In Fig. 9(a), the supply frequency can be estimated as shown in

$$\hat{f}_s = \frac{1}{4(T_c + T_d)}$$

The FIR low-pass filter (LPF) illustrated in Fig. 9(b) is adopted to the proposed frequency detector instead of a moving average (MA). The filtered output can be calculated as follows:

$$s[n] = \sum_{k=0}^{n} b_k s[n - k]$$

where $s[n]$ is the $n$th input signal, $s[n]$ is the $n$th filtered signal, and $b_k$ is the $k$th-order filter coefficient. This filter has seventh order and uses the Blackman window to enhance the sharpness of the filter at the cut-off frequency and to suppress sidelobe components. It can improve the transient dynamics and reduce the steady-state error of the frequency detector. Fig. 9(c) shows the block diagram of the single-phase SRF-PLL using the proposed frequency detector. The frequency detector calculates the fundamental frequency using (5) and it is added to the fuzzy value of $V_{ide}$ instead of the constant value to generate an accurate angle value with respect to frequency variations. In addition, the calculated frequency is used in the voltage generator.
which makes the orthogonal waveform to the supply voltage.

Fig. 9. Proposed single-phase PLL method: (a) frequency detection algorithm using zero crossing and APF, (b) block diagram of the seventh-order FIR filter, and (c) block diagram of the PLL using the proposed frequency estimator.

Because of the shorter update duration than the case of HD. In addition, QD with FIR

IV. ISOLATED BIDIRECTIONAL CLLC RESONANT CONVERTER

In this section, the design methodology of the power stages of the proposed bidirectional CLLC resonant converter will be discussed. The new control schemes are proposed to decide power flow directions and to regulate output voltage under bidirectional power flows. In addition, the dead-band control algorithm is proposed to smoothly change the power conversion direction only using output voltage information. This algorithm is based on a hysteresis control method to decide the power flow direction of bidirectional converters.

A. Gain Analysis Using the FHA model

The first harmonic approximation (FHA) model and operation mode in the proposed CLLC resonant converter are already introduced in [18]. Fig. 10 shows the equivalent circuit of the proposed full-bridge bidirectional CLLC resonant converter using the FHA method. The resonant network of the converter is composed of the series resonant capacitor $C_r = C_{r1} = C_{r2}$, the equivalent series resonant inductance $L_{eq}$, and the magnetizing inductance $L_m$. Since the turn ratio of the transformer is 1:1, it does not affect the circuit model and $L_{eq}$ equals the twice of the series resonant inductance $L_r = L_{r1} = L_{r2}$. The forward transfer function $H_r$ of the resonant network can be derived as shown in

$$H_{r}(s) = \frac{v_{r,FHA}(s)}{v_{in}(s)} = \frac{Z_{in}(s)}{Z_{in}(s) + R_{eq} + R_{op}}$$

$$Z_{in}(s) = \frac{1}{sL_{eq} + sR_{eq} + R_{eq}}$$

$$A(f_n) = f_n^{2}(1 + k) + K$$

$$B(f_n) = f_n^{2} - f_n^{2}(2 + K) + K$$

where $Z_{in}(s)$ is the input impedance of the resonant network, and $R_{eq}$ is the equivalent series resistance of the resonant circuit. The gain around the resonant frequency is determined by the ratio of $Z_{in}(s)$ and $R_{eq}$.
Fig. 10. FHA model of the proposed bidirectional CLLC resonant converter. Therefore, the gain of converter $H_r$ can be derived as follows:

$$f_{n}=\frac{f_{r}}{2\pi \sqrt{L_{R}C_{F}}}$$

$$Q=\frac{Z_{r}}{\sqrt{C_{F}L_{M}}}$$

where $A(fn)$ and $B(fn)$ are the function components as follows

$$f_s \leq f_r (12)$$

$$L_m \leq \frac{l_{dt}}{16C_f s_{\text{max}}} \leq l_{dt} (13)$$

**B. Dead-Band Control for Bidirectional Operation**

Fig. 13(a) shows the theoretical waveforms of the proposed dead-band control algorithm for the bidirectional CLLC resonant converter. When the load becomes negative, the output voltage of the converter will drastically increase because power is supplied to the output capacitor from two sides: the converter side and the load side. At this time, the converter is uncontrollable without changing the power conversion mode because of the negative power flow. If the output voltage reaches the positive dead-band voltage $+V_{\text{band}}$, the power conversion mode changes from the powering mode to the generating mode. In this generating mode, the converter transfers power from load to input side. Then, the output voltage will decrease to the reference voltage $V_{\text{ref}}$, which will be regulated by a pulse frequency modulation (PFM) controller. In the same manner, the power conversion mode can be changed from the generating mode to the powering mode. When the load becomes positive, the output voltage will decrease to the negative dead-band voltage $-V_{\text{band}}$. Then, the power conversion mode is changed and the output voltage will increase to $V_{\text{ref}}$. Fig. 11(b) shows the block diagram of the proposed digital controller. The dead-band controller can select the power conversion mode using the voltage gap $V_{\text{gap}}$, which is the voltage difference between the output voltage and the dead-band voltage. This dead-band controller generates the sign of the voltage error $V_{\text{err}}$ which is the voltage difference between the output voltage and the reference voltage. The fuzzy logic controller regulates the output voltage using $V_{\text{err}}$ and its sign. The switch control block generates PFM switching pulses using the calculated switching frequency generated by the fuzzy logic controller. Finally, the PFM switching pulses are assigned to the proper power switches considering the power conversion mode.

**V. FUZZY LOGIC CONTROLLER**

In FLC, basic control action is determined by a set of linguistic rules. These rules are determined by the system. Since the numerical variables are converted into linguistic variables, mathematical modeling of the system is not required in FC. The FLC comprises of three parts: Fuzzification, interference engine and defuzzification. The FC is characterized as i. seven fuzzy sets for each input and output. ii. Triangular membership functions for simplicity. iii. Fuzzification using continuous universe of discourse. iv. Implication using Mamdani’s ‘min’ operator. v. Defuzzification using the height method.

**Fuzzification:** Membership function values are assigned to the linguistic variables, using seven fuzzy subsets: NB (Negative Big), NM (Negative Medium),...
NS (Negative Small), ZE (Zero), PS (Positive Small), PM (Positive Medium), and PB (Positive Big). The partition of fuzzy subsets and the shape of membership function adapt the shape up to appropriate system. The value of input error and change in error are normalized by an input scaling factor.

**Table I Fuzzy Rules**

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<tr>
<th>Change in error</th>
<th>Error</th>
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In this system the input scaling factor has been designed such that input values are between -1 and +1. The triangular shape of the membership function of this arrangement presumes that for any particular $E(k)$ input there is only one dominant fuzzy subset.

The input error for the FLC is given as

$$E(k) = \frac{P_{ph(k)} - P_{ph(k-1)}}{V_{ph(k)} - V_{ph(k-1)}}$$

$$CE(k) = E(k) - E(k-1)$$

**Fig. 12. Fuzzy logic controller**

**Fig. 13. Membership functions**

**Inference Method:** Several composition methods such as Max–Min and Max-Dot have been proposed in the literature. In this paper Min method is used. The output membership function of each rule is given by the minimum operator and maximum operator. Table 1 shows rule base of the FLC.

**Defuzzification:** As a plant usually requires a non-fuzzy value of control, a defuzzification stage is needed. To compute the output of the FLC, „height” method is used and the FLC output modifies the control output. Further, the output of FLC controls the switch in the inverter. In UPQC, the active power, reactive power, terminal voltage of the line and capacitor voltage are required to be maintained. In order to control these parameters, they are sensed and compared with the reference values. To achieve this, the membership functions of FC are: error, change in error and output.

The set of FC rules are derived from

$$u=\alpha E + (1-\alpha)C$$

Where $\alpha$ is self-adjustable factor which can regulate the whole operation. $E$ is the error of the system, $C$ is the change in error and $u$ is the control variable. A large value of error $E$ indicates that given system is not in the balanced state. If the system is unbalanced, the controller should enlarge its control variables to balance the system as early as possible. One the other hand, small value of the error $E$ indicates that the system is near to balanced state.
VI. CONCLUSION

The isolated bidirectional ac–dc converter is proposed for the 380-V dc power distribution system to control the bidirectional power flow and to improve its power conversion efficiency using fuzzy logic controller. In order to improve the reverse recovery problem, the high-side switches of the ac–dc rectifier employ IGBTs without anti parallel diodes and SiC diodes. In addition, the low-side switches are composed of two MOSFETs to reduce the conduction loss in the rectification mode. For comparison with the conventional IGBT switches, the total conduction losses of the rectifier’s switches are calculated in the rectification mode. The simple and intuitive frequency detection method for the single-phase SRF-PLL is also proposed using the filter compensator, fast QD, and FIR filter to improve the robustness and accuracy of the PLL performance under fundamental frequency variations. The proposed PLL system shows lower detection fluctuation and faster transient response than the conventional techniques. Finally, the proposed LLC resonant converter can operate under the ZVS for the primary switches and the soft commutation for the output rectifiers. The soft-switching condition of the converter is derived to obtain the design methodology of the resonant network. Gain properties are also analyzed to avoid gain reduction and no monotonic gain curve under high-load conditions. In addition, the dead-band and switch transition control algorithms are proposed to smoothly change the power flow direction in the converter.
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