Unidirectional DC/DC modular multilevel converter for offshore windfarm with the control strategy based on stationary frame

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Abstract: This paper presents the design and control of an advanced unidirectional DC/DC modular multilevel converter (MMC) that facilitate the integration of off-shore windfarms with the high-voltage direct current (HVDC) transmission system. The proposed converter deploys a single-phase MMC inverter coupled with series-connected rectifier modules through multiple medium frequency transformers. Unlike the conventional dq control method, which involves multiple transformations, this paper also proposes a simple control strategy that directly acts on the AC output of the MMC, under the stationary reference frame.

1 Introduction

High-voltage direct current (HVDC) transmission system is a well-established and proven technology for delivering large-scale energy over a long distance with less power losses and lower reactive power requirements [1]. Large-scale offshore wind energy is increasingly growing and the interconnection between multiple farms becomes more challenging. Medium-voltage DC collection networks are a promising technology for such integration aiming to eliminate the extra conversion stages and improve the system reliability [2]. High-voltage high-power DC/DC converters are the key enabler for the DC grid. Various converter topologies have been investigated and reported in the literature, which can be broadly classified as combined (consisting of multiple converter modules) and modular multilevel topologies [3].

Most notably, dual active bridge (DAB) converter has received a great attention from the research community due to its distinctive features, such as galvanic isolation, bidirectional power flow, and ability to operate with high switching frequency [4]. However, the high-voltage and high-power requirements for the DC/DC converter-based HVDC systems, necessitate series and/or parallel combinations at both, the power semiconductor devices and converter modules levels [5]. Furthermore, as the requirement for offshore DC collection point is to deliver a high-voltage, facilitating the connection with HVDC transmission system, the input parallel output series (IPOS) configuration is commonly preferred [6], which is retained for this work as well. There are several papers investigated IPOS combined converter as DC collection point for the HVDC system [7–11]. However, for such a converter, the full soft switching operation can only be achieved with a limited load and input voltage range, which substantially limits the efficiency and the performance of the converter due to the increased switching losses and electromagnetic interference [8].

To address this problem, an external large resonant inductor is usually connected in series with the transformer to extend the soft switching range, but the large inductance has a detrimental effect on the performance of the converter since it results in increased duty cycle losses, as well as a severe voltage ringing due to the resonance between the inductance and the junction capacitance in the converter [9]. The concept of using a saturation inductor instead of linear inductor has been discussed in [11], which effectively extends the soft switching range with lower conduction losses and without a significant duty cycle loss. However, a large core is required for thermal dissipation, limiting the whole system power density and large-scale applications.

The research here alleviates the above-mentioned issues by proposing a modular DC/DC converter, employing MMC at the primary side of a medium-frequency transformer. The DC voltage is collected at the secondary side through series-connected diode-bridge rectifier modules. The proposed system is highly modular and considerably reduces the switching losses. An enhanced control method is also proposed to control the operation of the converter.

The rest of the paper is organised as follows: Section 2 describes the circuit configuration of the proposed DC/DC converter and its operating principle. The analysis of power balance of MMC at the primary side is presented in Section 3. A simple control strategy based on the stationary reference frame is derived in Section 4. Section 5 illustrates simulation results. Finally, the work is concluded in Section 6.

2 Proposed DC/DC converter-based system

Fig. 1 shows a simplified schematic diagram of the proposed converter which can be functioned as DC collection point, where a single-phase (two-leg) MMC inverter producing a controllable AC voltage is connected at the primary side of multiple medium frequency (400 Hz) transformers. The DC output voltage is obtained through series-connected full-bridge rectifier modules at the multi-winding secondary side of the transformers. It is worth noting that the design is fully modular at both sides and can be easily expanded as required by simply adding more modules.

3 Mathematical model of the proposed converter

Fig. 2 shows the equivalent circuit of one-leg (phase A) of the MMC, where \( V_{dc, in} \) and \( I_{dc, in} \) are the converter’s DC input voltage and current, respectively. \( V_{ap} \) and \( V_{an} \) are the upper and lower arm voltages of the cascade submodules of Phase A leg, respectively. \( I_{ap} \) and \( I_{an} \) are the current of the upper and lower arms, respectively. \( E_o \) is the equivalent output phase voltage as shown in Fig. 2(b) and \( V_o \) is output AC voltage, respectively. \( I_{cir} \) and \( I_o \) are circulating current and output AC current, respectively.

From Fig. 2, the upper and lower arm currents of Phase A leg can be expressed as:

\[
I_{ap} = I_o/2 + I_{cir} \\
I_{an} = -I_o/2 + I_{cir}
\]

(1)

(2)

where the circulating current, \( I_{cir} \), is flowing through both the upper and lower arms.
Fig. 1 Typical schematic diagram of off-shore HVDC transmission system using the proposed DC/DC converter functioned as DC collection point

Fig. 2 Schematic diagram of (a) one-leg MMC, (b) and its equivalent circuit

It should be noted that the circulating current has no effect on the output phase current and can be expressed as:

\[ I_{i.c.} = (I_{dp} + I_{dn}) / 2 \]  

(3)

With reference to (1) and (2), the equation of output AC current \( I_a \) can be expressed in terms of upper and lower arm currents as:

\[ I_a = I_{dp} - I_{dn} \]  

(4)

Considering \( n \) as the neutral point, applying the Kirchhoff Voltage Laws (KVL) for the schematic diagram of one-leg MMC as shown in Fig. 2a, therefore, the upper and lower voltages can be derived as:

\[ V_{dp} = \frac{V_{dc}}{2} - V_a - I_{arm} \frac{dI_{dp}}{dt} \]  

(5)

\[ V_{dn} = \frac{V_{dc}}{2} + V_a - I_{arm} \frac{dI_{dn}}{dt} \]  

(6)

Combining (5) and (6), the output phase voltage \( V_a \) can be expressed as:

\[ V_a = \frac{1}{2} (V_{dn} - V_{dp}) - \frac{1}{2} I_{arm} \frac{d(I_{dn} - I_{dp})}{dt} \]  

(7)

Substituting (4) into (7), the equivalent output phase voltage \( E_a \) can be given by:

\[ E_a = \frac{1}{2} (V_{dn} - V_{dp}) = V_a + \frac{1}{2} I_{arm} \frac{dI_a}{dt} \]  

(8)

Therefore, the mathematical model of the one-leg MMC can be derived by rearranging (8) as:

\[ \frac{1}{2} I_{arm} \frac{dI_a}{dt} = E_a - V_a \]  

(9)

According to (9), the equivalent circuit of Phase A can be expressed as Fig. 2b.

Similarly, the mathematical model of the second leg of MMC (i.e. phase B) can be given by:

\[ \frac{1}{2} I_{arm} \frac{dI_b}{dt} = E_b - V_b \]  

(10)

where \( E_b \) and \( V_b \) are the equivalent output phase voltage and output AC voltage of Phase B, respectively and \( I_b \) is the output AC current of Phase B.

Combining (9) and (10), the mathematical model of single phase (two-leg) MMC can be expressed as:

\[ \frac{1}{2} I_{arm} \frac{dI_a}{dt} - \frac{1}{2} I_{arm} \frac{dI_b}{dt} = (E_a - E_b) - (V_a - V_b) \]  

(11)

For simplicity, let:

\[ V_{P0} = E_a - E_b \]  

(12)

\[ V_{ab} = V_a - V_b \]  

(13)

Hence, (11) can be re-written as:

\[ \frac{1}{2} I_{arm} \frac{dI_a}{dt} = V_{P0} - V_{ab} \]  

(14)

For a single-phase (two-leg) MMC, the relationship between the output current of phase A and B can be expressed as:

\[ I_P = I_a = -I_b \]  

(15)

where \( I_P \) is the transformer primary current of the proposed converter.

Substituting (15) into (14), yields:

\[ I_{arm} \frac{dI_a}{dt} = V_{P0} - V_{ab} \]  

(16)

where the \( V_{P0} \) and \( V_{ab} \) can be considered as the equivalent primary voltage and primary terminal voltage of the transformer, respectively. \( I_P \) is transformer primary voltage. \( I_{arm} \) is arm inductance.

From Fig. 1, the primary side of the transformers are connected in parallel, which are equivalent to a parallel connection of voltage sources (\( V_{P0} \)). Similarly, the secondary side of the transformer is made of a combination of individual and isolated modules. This can be regarded as a series connection of voltage sources (\( V_{S1}, V_{S2}, \ldots, V_{SNs} \)). Therefore, the total equivalent voltage at the secondary side of the transformer \( V_s \), can be expressed as:

\[ V_s = V_{S1} + V_{S2} + \cdots + V_{SNs} \]  

(17)

where \( N_s \) is the number of rectifier modules at the secondary. If the equivalent primary-to-secondary winding turns ratio is \( R_t \) and the turns ratio of primary to each individual secondary winding is \( T_r \), the equivalent secondary voltage \( V_s \) when it is referred to the primary side can then be given by:

\[ V_{ab} = V_s R_t = V_{SNs} T_r N_s + V_{SNs} T_r N_s + \cdots + V_{SNs} T_r N_s \]  

(18)
Fig. 3 Equivalent circuit of
(a) proposed converter referred to primary side, (b) and its simplify equivalent circuit

Substituting (16) into (18), and with the transformer's leakage inductance $L_k$ referred to the primary side, the primary referred equivalent circuit of the proposed converter can be expressed by (19) and schematically represented by Fig. 3.

$$L_{	ext{air}} + L_k \frac{dI_p}{dt} = V_{po} - V_s R_t$$

(19)

4 Power transfer characteristics of the proposed converter

4.1 Voltage and current key-waveforms of the proposed converter

To simplify the analysis and to facilitate easy understanding, the following assumptions are considered for the output voltage waveform of the MMC. (1) The submodule capacitor voltages are well balanced and ripple-free. (2) The converter is operating with a unity modulation index. (3) With high number of submodules, resulting in small steps in the output waveform, which then can be approximated as trapezoidal waveform as illustrated in Fig. 4. It is worth noting also that, in this work, the well-known carrier-phase-shift pulse-width modulation (CPS-PWM) technique is applied. The detailed analysis and the principle of operation of CPS-PWM have been well documented and readily available from [12]; therefore, no further details will be presented here.

Fig. 4 shows typical voltage of primary-referred equivalent circuit of the proposed converter, where the power is always transferred from $V_{dc\text{-in}}$ to $V_{dc\text{-out}}$ (i.e. unidirectional). It should be noted that $\theta_{\text{stair}}$ in Fig. 4 is the interval of the voltage rising or falling transition of the $V_{po}$, $V_s$ is the secondary equivalent voltage which is equal to the sum of $V_s + V_{s1} + \cdots + V_{sN_s}$ and $\varphi$ is the phase-shift angle by which $V_s$ lags $V_{po}$.

4.2 Derivation of the proposed converter's output power

It should be noted that in the succeeding analysis, the following assumptions are considered: (1) owing to the waveform symmetry, only half-cycle of the AC output waveform is considered as shown in Fig. 4; and (2) for simplicity, the peak value of $V_{po}$ and $V_s$ are equal to $V_{dc\text{-in}}$ and $V_{dc\text{-out}}$, respectively. (i.e. ignoring any voltage drop across the circuit components). According to Fig. 4, the output power can be derived in the following operational intervals:

Interval 1 ($0 \leq \theta < \theta_{\text{stair}}/2$):

As it can be noticed from Fig. 4, during this interval, $V_s(\theta)$ is equal to the $-V_{dc\text{-out}}$ and its primary referred voltage is $V_{dc\text{-out}} R_t$ Therefore, $V_{po}(\theta)$ and $I_p(\theta)$ can be expressed as:

$$V_{po}(\theta) = \frac{2V_{dc\text{-in}}}{\theta_{\text{stair}}} \theta$$

(20)

$$I_p(\theta) = I_p(0) + \frac{1}{L} \int_0^\theta (V_{po}(\theta) - R_s V_s(\theta)) d\theta$$

(21)

Replacing $V_s(\theta)$ by $-V_{dc\text{-out}}$ and substituting (20) into (21), yields:

$$I_p(\theta) = I_p(0) + \frac{1}{L} \int_0^{\theta_{\text{stair}}/2} (V_{dc\text{-in}} - V_{dc\text{-out}} R_t) d\theta$$

(22)

According to (22), at $\theta = (\theta_{\text{stair}}/2)$, one can get:

$$I_p(\theta_{\text{stair}}/2) = I_p(0) + \left(\frac{\theta_{\text{stair}}}{2L}\right)(V_{dc\text{-in}} - V_{dc\text{-out}} R_t)$$

(23)

According to (20) and (22), the output energy during this interval can be obtained by:

$$E_1 = \int_0^{\theta_{\text{stair}}/2} V_s(\theta) \times I_p(\theta) d\theta$$

$$= -V_{dc\text{-out}} \times \theta_{\text{stair}} / 2 \times \left(\frac{3V_{dc\text{-in}} + 3V_{dc\text{-out}} R_t + 3R_s V_s(0) + 12LR_s(0) I_p(0)/24}{L}\right)$$

(24)

Interval 2 ($\theta_{\text{stair}}/2 \leq \theta < \varphi$):

During this interval, $V_s(\theta)$ remains as $-V_{dc\text{-out}}$, $V_{po}(\theta)$ is equal to the $V_{dc\text{-in}}$, and $I_p(\theta)$ can be expressed as:

$$I_p(\theta) = I_p(0) + \frac{1}{L} \int_0^{\theta_{\text{stair}}/2} (V_{po}(\theta) - R_s V_s(\theta)) (\theta - \theta_{\text{stair}}/2) d\theta$$

(25)

Substituting $V_s(\theta) = -V_{dc\text{-out}}$, $V_{po}(\theta) = V_{dc\text{-in}}$ and (25) into $I_p(\theta)$ yields

$$I_p(\theta) = I_p(0) + \frac{1}{L} \left\{V_{dc\text{-in}} + V_{dc\text{-out}} R_t - \frac{V_{dc\text{-out}} \theta_{\text{stair}}}{4}\right\}$$

(26)

From (26), at $\theta = \varphi$, one can get:

$$I_p(\varphi) = I_p(0) + \frac{1}{L} \left\{V_{dc\text{-in}} + V_{dc\text{-out}} R_t \varphi - \frac{V_{dc\text{-out}} \theta_{\text{stair}}}{4}\right\}$$

(27)

Similarly, the transferred energy during this interval is given by: (see (28))

Interval 3 ($\varphi \leq \theta < \pi - (\theta_{\text{stair}}/2)$):

During this interval, $V_s(\theta)$ is equal to $V_{dc\text{-out}}$, $V_{po}(\theta)$ remains as $V_{dc\text{-in}}$, and $I_p(\theta)$ can be expressed as:

$$\frac{E_3}{8L} = \frac{V_{dc\text{-in}} (\theta_{\text{stair}} - 2\varphi)(V_{dc\text{-out}} \theta_{\text{stair}} R_t^2 + 2V_{dc\text{-out}} \varphi R_t^2 + 2R_s V_{dc\text{-in}} \varphi + 4LR_s(0) I_p(0))}{8L}$$

(28)

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Substituting $V_p(\theta) = V_{\text{dc, in}}$, $V_{\theta}(\theta) = V_{\text{dc, out}}$, and (27) into (29) yields

$$I_p(\theta) = I_p(\theta) + \frac{1}{L} \left[ V_{\text{dc, in}} - V_{\text{dc, out}} R_i \right] \theta + 2 V_{\text{dc, out}} R_i \theta - \frac{V_{\text{dc, out}} \theta_{\text{stair}}}{4} \tag{30}$$

From (30), at $\theta = \pi - (\theta_{\text{stair}}/2)$, one can get:

$$I_p(\pi - \frac{\theta_{\text{stair}}}{2}) = I_p(0) + \frac{1}{L} \left[ V_{\text{dc, in}} - V_{\text{dc, out}} R_i \right] \left( \pi - \frac{\theta_{\text{stair}}}{2} \right) + 2 V_{\text{dc, out}} R_i \theta - \frac{V_{\text{dc, in}} \theta_{\text{stair}}}{4} \tag{31}$$

Similarly, the output energy during this interval can be expressed as:

$$E_i = \int_{\pi - \theta_{\text{stair}}/2}^{\pi} V_p(\theta) \times I_p(\theta) \, d\theta = -\frac{1}{2L} \int_{0}^{\theta_{\text{stair}}/2} V_{\text{dc, out}} \theta_{\text{stair}} - 2 \pi + 2 \phi \, d\theta$$

$$\left( V_{\text{dc, out}} \theta_{\text{stair}} R_i - 2 \pi V_{\text{dc, out}} R_i + 6 V_{\text{dc, out}} \theta_{\text{stair}} R_i - 2 R_i \times V_{\text{dc, out}} \theta_{\text{stair}} + 2 R_i V_{\text{dc, in}} + 4 L R_i \right)$$

Interval 4 ($\pi - (\theta_{\text{stair}}/2) \leq \theta < \pi$):

During this interval, $V_p(\theta)$ remains as $V_{\text{dc, out}}$. The primary voltage, $V_{\text{dc, out}}$ and the primary current, $I_p(\theta)$ can be expressed as:

$$V_{\text{dc, in}} \theta_{\text{stair}} (2 \pi - 2 \pi) \tag{33}$$

$$I_p(\theta) = I_p(\theta) + \frac{1}{L} \int_{\pi - \theta_{\text{stair}}/2}^{\pi} V_{\text{dc, out}} \theta_{\text{stair}} - 2 \pi V_{\text{dc, out}} R_i \, d\theta$$

$$I_p(\pi - \frac{\theta_{\text{stair}}}{2}) = I_p(0) + \frac{1}{L} \left[ V_{\text{dc, in}} - V_{\text{dc, out}} R_i \right] \left( \pi - \frac{\theta_{\text{stair}}}{2} \right) + 2 V_{\text{dc, out}} R_i \theta - \frac{V_{\text{dc, in}} \theta_{\text{stair}}}{4} \tag{35}$$

From (35), with $\theta = \pi$, one can write the primary current as:

$$I_p(\pi) = I_p(0) + \frac{1}{L} \left[ \pi V_{\text{dc, in}} - 2 V_{\text{dc, out}} \theta_{\text{stair}} \right] + 2 V_{\text{dc, out}} \theta_{\text{stair}} \tag{36}$$

Hence, from (33) and (35), the output energy during this interval can be obtained by (37) below.

$$E_i = \int_{\pi - \theta_{\text{stair}}/2}^{\pi} V_p(\theta) \times I_p(\theta) \, d\theta$$

$$= \frac{1}{2L} \left[ \left( V_{\text{dc, in}} \theta_{\text{stair}} R_i - 2 \pi V_{\text{dc, in}} R_i + 6 V_{\text{dc, in}} \theta_{\text{stair}} R_i - 2 R_i \times V_{\text{dc, in}} \theta_{\text{stair}} + 2 R_i V_{\text{dc, in}} + 4 L R_i \right) \right]$$

Therefore, from (24), (28), (32), and (37), the output power of the proposed converter can be calculated as:

$$P_{\text{out}} = E_i + E_2 + E_3 + E_4 \tag{38}$$

Due to half-cycle symmetry, $I_p(0) = -I_p(\pi)$ Therefore, according to (36), the initial current $I_p(0)$ can be calculated as:

$$I_p(0) = 2 \pi V_{\text{dc, out}} R_i - 4 V_{\text{dc, out}} \theta_{\text{stair}} R_i + V_{\text{dc, out}} \theta_{\text{stair}} - 2 \pi V_{\text{dc, in}} \tag{39}$$

Letting $G = (V_{\text{dc, out}} R_i)/V_{\text{dc, in}}$ in (38) and (39), the output power of the proposed converter at any phase shift angle, $\phi$, can be expressed by:

$$P_{\text{out}}(\phi) = -G V_{\text{dc, in}} (2 \pi \phi^2 - 12 \pi \phi) \tag{40}$$

where $G$ is defined with the primary-referred DC voltage gain of the proposed converter, often referred to as the DC conversion ratio.

### 4.3 Analysis of the proposed converter’s output power

Similar to the conventional single active bridge (SAB) DC/DC converter [13], based on the different phase-shift angle, the proposed converter has two operating modes: (1) $0 \leq \phi < (\theta_{\text{stair}}/2)$ and (2) $(\theta_{\text{stair}}/2) \leq \phi < (\pi/2)$. Therefore, the corresponding $I_p$ in the different modes can also be obtained. Then, the output power in the different modes can be derived as:

$$P_{\text{out, norm}} = \begin{cases} \frac{M^2 \phi (3 \theta_{\text{stair}} - 6 \pi \theta_{\text{stair}} + 4 \phi^2)}{2} & 0 \leq \phi < \frac{\theta_{\text{stair}}}{2} \\ \frac{M^2 (\theta_{\text{stair}} + 12 \phi^2 - 12 \pi \phi)}{2} & \frac{\theta_{\text{stair}}}{2} \leq \phi \leq \frac{\pi}{2} \end{cases} \tag{41}$$

where modulation index $M \in [0, 1]$ is introduced in order to get the generalised equation of the output power.

Fig. 5 illustrates the variation of the normalised power of (41) with respect to phase-shift angle $\phi$. It is clear that the power transfer capability of the proposed converter is influenced by the modulation index $M$, $\theta_{\text{stair}}$, and the phase-shift angle $\phi$, where the highest power is achieved when the modulation index is unity. On the other hand, increasing $\theta_{\text{stair}}$ decreases the power transfer capability of the proposed converter.

### 5 Control strategy of the proposed converter

Considering one-leg MMC as an example, the implementation of the proposed closed-loop control is illustrated in Fig. 6. It is realised by multiplying the transfer functions of the conventional proportional integral (PI) regulators with the sin and cos reference signals. Equation (42) describes the implementation of the proposed control method, which is schematically depicted in Fig. 6.

![Fig. 5: Normalised output power versus phase shift angle $\phi$ of the proposed converter](http://creativecommons.org/licenses/by/3.0/)
Fig. 6 Block diagram of implementation of the close loop control

Table 1 Parameters of the simulation

| Parameter                          | Value          |
|-----------------------------------|----------------|
| rated power                       | 80 MW          |
| input DC voltage                  | 100 kV         |
| output DC voltage                 | 200 kV         |
| submodule numbers of MMC per arm  | 15             |
| submodule numbers of combined converter | 2          |
| transformer ratio                  | 2:3:3          |
| submodule capacitor of MMC         | 2.2 mF         |
| inductance of per arm              | 1 mH           |
| output capacitor                  | 3 mF           |
| output inductor                   | 1 mF           |
| switching frequency               | 2000 Hz        |
| AC fundamental frequency          | 400 Hz         |

\[
V_{ac}^{\text{ref}}(t) = \left[ e_{ac}(t) \times \sin(\omega t) \right] * h_{\text{ac}}(t) \times \sin(\omega t)
+ \left[ e_{ac}(t) \times \cos(\omega t) \right] * h_{\text{ac}}(t) \times \cos(\omega t) \tag{42}
\]

where \(e_{ac}(t)\) is the input AC error signal, \(h_{\text{ac}}(t)\) represents the unit impulse response under time domain of PI regulator, and * denotes the convolution product. The output of this control loop \(V_{ac}^{\text{ref}}\) is used as a modulating signal to drive the power switches of phase A.

Actually, after multiplying with reference signal, sin and cos, the error signal \(e_{ac}(t)\) is converted into a DC and AC components with two times the fundamental frequency. This is then fed into the PI regulators, which perform the integration to get the steady-state error and also work as a low-pass filter to extract out the DC signal to achieve a zero steady-state error in the stationary reference frame. Applying Laplace transform to (43) yields

\[
V_{ac}^{\text{ref}}(s) = \frac{1}{\pi} \left( H_{\text{ac}}(s + j\omega) + H_{\text{ac}}(s - j\omega) \right) e_{ac}(s) \tag{43}
\]

For a conventional PI regulator, the Laplace transform of a unity impulse response can be described by the following equation:

\[
H_{\text{ac}}(s) = k_p + \frac{k_i}{s} \tag{44}
\]

By substituting (66) into (65), the transfer function of the control system with derived generalised integrator can be given as:

\[
V_{ac}^{\text{ref}}(s) = \left( k_p + \frac{k_i \times s}{s^2 + \omega^2} \right) e_{ac}(s) \tag{45}
\]

where \(k_p\) is the proportional constant which is used to improve the transient response of the control system, \(k_i\) is the integral constant and \(\omega\) is the resonant frequency of the derived integrator. Therefore, the infinity gain can be achieved when \(s = j\omega\)

6 Simulation results

A simulation model of the proposed converter rated at 80 MW/200 kV is developed with the tabulated parameters in Table 1 using MATLAB/SIMULINK to validate the feasibility and the effectiveness of the proposed system and its control performance. Here, the MMC is constructed with fifteen half-bridge SM per arm, and there are two series-connected diode-bridge rectifiers at the secondary side of the transformer. Furthermore, the medium-frequency transformer with turn ratios of 2:3:3 is chosen; however, different turns’ ratio can be simply considered for different stepping gain.

The steady-state voltage and current of the proposed converter, operating at 400 Hz, is depicted in Fig. 7.

The output DC voltage of the proposed converter is shown in Fig. 8 and with the given transformer turns ratio here (i.e. \(W_p:W_{s1}:W_{s2} = 2:3:3\), where \(W_p\), \(W_{s1}\), and \(W_{s2}\) are the primary winding, the corresponding secondary winding-one and winding-two, respectively), the average output DC voltage is maintained around 200 kV.

The performance of the proposed control strategy is further investigated and confirmed with a step change in the output load (i.e. 500 to 600 Ω). Fig. 9 shows the dynamic response of the controller when the load changed at \(t = 2.5\) s, causing the output current \(I_{dc2}\) to decrease from 400 to 333 A; however, the output voltage is perfectly maintained constant, which confirms the effectiveness of the proposed control system.

7 Conclusion

A modular unidirectional DC/DC converter-based DC collection point is presented here. The proposed converter utilised the state-
of-the-art MMC at the primary side of a medium-frequency transformer and cascaded diode-bridge rectifier modules at the secondary side. The converter design features modularity, expandability, redundancy, galvanic isolation, and lower voltage and current stress on power devices. Detailed design analysis is presented and the parameters that affect the operation of the converter are defined and thoroughly discussed. The presented control strategy abolishes all complexity associated with the transformation and inverse transformation required for the conventional d-q synchronous reference frame methods. The derived generalised controller directly acts on the AC single of the primary side, avoiding complex transformation and providing robustness against system variations. The performance of the proposed converter and its control strategy is validated through various simulation results.

8 References

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