Transient Behaviour of Solid State Modulators with Matrix Transformers

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***Abstract*—Solid state modulators based on pulse transformer offer the advantage, that with the turns ratio of the transformer, the primary voltage ideally could be adapted to the available switch technology and a series connection of switches could be avoided. For increasing the power level several semiconductor switches must be connected in parallel and a balancing between the different switches must be guaranteed. There, the Matrix Transformer concept, which is based on multiple primary and/or secondary windings as well as on flux adding, offers superior performance with respect to the achievable rise times. However, the influence of the parasitic elements on the voltage and current distribution is quite involved.**

In this paper the influence of the parasitic elements of the Matrix Transformer on the current balancing and winding voltages is investigated based on reluctance models and the inherent current balancing of the Matrix Transformer for windings mounted on different cores is explained. Furthermore, the influence of parasitic load/transformer capacitances on the turn-off transient is discussed in detail.

I. Introduction

In many pulsed power applications, as for example in the medial or in the accelerator area, rectangular pulses with a flat top, a fast rise time, and variable pulse width are required. For generating such pulses with variable pulse width, basically three different modulator types: direct switch modulators [1], multi-cell-type generators (e.g. Marx-generator) [2], and transformer based modulators [3] as well as a combination of them, can be used.

In all of these modulator types, the achievable rise and fall times of the pulses are mainly determined by the parasitic capacitances and inductances in the power circuit, consisting of for example, the capacitor bank, the switches, the interconnections, and the load. Due to the insulation requirements, the same minimal distances are required in the design of any type of modulator. Therefore, in all modulator types the achievable rise and fall times are in the same order of magnitude. Furthermore, the voltage droop of the pulse top is relatively independent of the modulator type, since it is determined mainly by the amount of stored energy and/or by an existing droop compensation. Consequently, no modulator concept has fundamental advantages with respect to the achievable transients and the voltage droop.

However, modulators based on pulse transformers offer an additional degree of freedom – the turns ratio – which enables the adaption of the primary voltage on the available switch technologies, so that advantageously also standard semiconductor switches, used for example in traction applications, can be employed in the solid state modulator. Moreover, a series connection of switching devices with its critical dynamic voltage balancing could be avoided.

For increasing the pulsed power of a transformer based modulator system, the switches on the primary side of the transformer could be connected in parallel, what is basically simpler than a series connection. The current balancing between the parallel connected switches could be for example relatively simple achieved by scheduling the gate pulses with an active gate control, as shown in [4]. Moreover, for some special transformer concepts, e.g. Matrix/Split Core transformer, an inherent current balancing between parallel connected switches is given, as will be explained in this paper. This simplifies the design of the modulator system further. Additionally, the Matrix Transformer leads to a reduction of the leakage inductance, which results in an improved pulse performance.

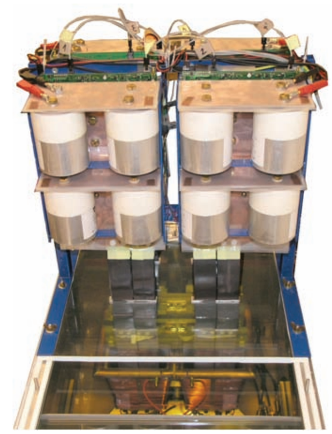

Fig. 1: Photo of a 20MW solid state modulator with Matrix (Split Core) Transformer and the specifications given in Table I.

A second advantage of the transformer based modulators is the reduction of the switching losses and the overvoltage at turn-off, due to the parasitic capacitance of the pulse transformer and the load. Additionally, this effect could also result in faster turn-off times, as will be shown in this paper.

So far, the inherent current sharing of the Matrix Transformer and the reduction of the turn-off losses for modulators with pulse transformer have not been explained in detail. Therefore, in Section II the influence of the parasitic transformer/load capacitances on the turn-off transient are discussed. Thereafter, the focus is put on transformers with multiple windings and/or cores, which in general are called Matrix Transformer. First, the basic operating principle of the Matrix Transformer is shortly explained based on reluctance models and compared to other transformer designs in

TABLE I: Specification of the $20\,\mathrm{MW}$, $5\,\mu\mathrm{s}$ pulse modulator with four parallel connected IGBT modules and Matrix Transformer.

DC Link Voltage V_{DC}	$1\,\mathrm{kV}$
Output Voltage V_{out}	$170\,\mathrm{kV}$
Pulse Duration T_{pulse}	$5\,\mu\mathrm{s}$
Rise Time T_{rise}	$< 500\,\mathrm{ns}$
Output Power P_{out}	$20\,\mathrm{MW}$
Repetition Frequency f_{rep}	$200\,\mathrm{Hz}$
Conversion Ratio	$1:170$

Section III. Additionally, the inherent current balancing between windings mounted on different cores is explained. Thereafter, the transient flux and voltage distribution in case of synchronous and asynchronous operation of the primary switches is explained and measurement results for the solid state modulator shown in Fig. 1 with the specifications given in Table I are shown in Section IV.

II. Inherent Capacitive Turn Off "Snubber"

Independent of the chosen modulator type, the series inductance L_{Com} in the commutation path, consisting for example of the parasitic inductances in the switches, the interconnections, and the capacitor bank, results in overvoltages during turn-off. In Fig. 2a) the turn-on and turn-off behaviour of a 1.7kV-3.6kA IGBT module connected to a resistive load is shown. There, clearly the resulting overvoltage at turn-off of approximately 200V could be seen.

In order to limit the overvoltages to safe levels, usually a reduction of the di/dt and therewith an increase of the pulse fall time is required. However, with multi-stage gate drives using different gate resistors during turn-off, or gate drives with zener clamping [4] the pulse fall-time could be reduced.

In transformer based power modulators, however, the distributed capacitance of the pulse transformer helps to limit the overvoltage and to reduce the switching losses also without multi-stage gate drives, as will be explained based on the circuit diagram shown in Fig. 3a). There, a simplified equivalent circuit of a single-switch transformer-based power modulator with capacitor bank, semiconductor switch, simple transformer model and resistive load is shown. L_{Com} is the series inductance in the commutation path, comprising the parasitic inductance of the capacitor bank, the switch, and the interconnection between capacitor bank/switch. L_σ represents the leakage inductance, L_{mag} the magnetizing inductance of the transformer and C_d the distributed capacitance of the pulse transformer referred to the primary side.

During the pulse, capacitance C_d is approximately charged to the input voltage V_{DC} (cf. Fig. 3). Assuming a relatively large capacitance C_d, where V_{C_d} changes significantly slower than V_{CE}, the voltage across the parasitic inductors L_{Com} and L_σ becomes negative as soon as the switch S_1 is turned-off and the voltage V_{CE} starts to rise as could be seen in the simulated waveforms given in Fig. 3b). Therefore, the current in the inductors as well in the switch S_1 immediately starts to decrease, which results in lower switching losses of the IGBT module compared to the case without capacitor C_d.

As soon as the current in the inductor L_{Com} reaches zero also the voltage across the inductor becomes zero. At this point of time, V_{CE} drops relatively fast to the value $V_{DC} - V_{C_d}$, as could be seen in Fig. 2b), where the negative voltage drop (oscillation)

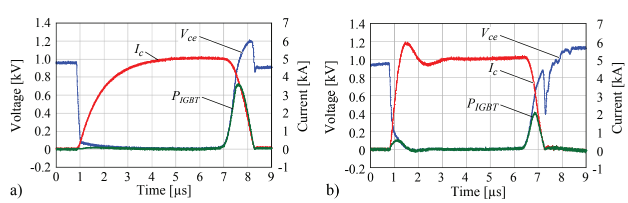

Fig. 2: Measured collector-emitter voltage V_{CE}, collector current I_C and switching losses P_{IGBT} of a 1.7kV/3.6kA IGBT with a) resistive and b) capacitive load.

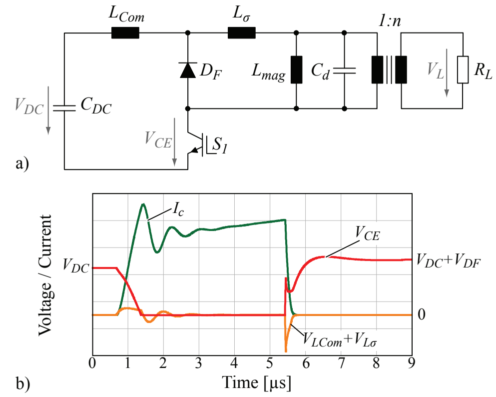

Fig. 3: a) Equivalent circuit of the single-switch transformer-based power modulator with transformer parasitics, as leakage inductance L_σ, magnetizing inductance L_{mag}, and distributed capacitance C_d. Assuming a purely resistive load, i.e. $C_d = 0$, the series inductance L_{Com} in the commutation path causes overvoltages during turn-off. b) Simulated collector current I_C, collector emitter voltage V_{CE} and voltage across L_{Com} and L_σ for $C_d > 0$, i.e. soft turn-off.

in the rising edge of V_{CE} occurs. At this time also the voltage of capacitor C_d will be smaller than the input voltage V_{DC}, since on the one hand C_d is discharged by the magnetizing current and on the other hand by the difference of the load current and i_{L_σ}.

After the current in the inductor L_{Com} reached zero, the rise time of the voltage $V_{CE} = V_{DC} - V_{Cd}$ is mainly determined by the voltage across the capacitance C_d, which is further discharged by the magnetising current and the load current. Consequently, no overvoltage can be caused by the series inductance L_{Com}. The overvoltage in Fig. 2b) results from the forward bias of the freewheeling path. Therefore, the IGBT can be turned off much faster, which allows a significant reduction of the turn-off losses.

However, for large values of C_d, the dv/dt could be limited by C_d, so that the rise times increases. In pulsed power systems with loads having a relatively large parasitic capacitance, as for example klystrons, similar effects can be observed, even without transformers.

III. Matrix Transformers

In the previous section the influence of the distributed capacitance C_d on the turn-off transient and the related reduction of the losses/pulse rise time has been discussed. This effect is relatively independent of the transformer configuration. The configuration, however, strongly influences the transient voltage and current distribution in the solid state modulator, as will be explained in the following.

In the most simplest case, the pulse transformer of the modulator has just one primary and one secondary winding wound around one core. Considering the pulse specifications given in Table I, i.e. $V_{out} = 170kV$ and $P_{out} = 20MW$, in a first step of the design process, the turns ratio must be chosen.

The turns ratio is strongly related to the applied switching technology due to the operating/blocking voltage. At the moment, the highest power rating of a single semiconductor switch capable to achieve the required rise time of smaller than 500ns are standard 1.7kV IGBT modules. These IGBTs are available for pulse currents of approximately 7.2kA, resulting in a pulsed power of ≈7MW per switch, if an operating voltage of 1kV is assumed. With these high power modules, the achievable rise/fall times are mainly limited by the housing designed for applications with low switching speeds as for example traction or high power drives [5].

Based on the mentioned maximum pulsed power ratings of the 1.7kV IGBT modules, a series or parallel connection of four IGBT modules (≈4x7MW) would be required for the modulator system with the specifications in Table I, if additional safety margins and the requirement to handle the magnetizing current are considered.

In general, with a series connection the static and dynamic balancing of the voltage distribution between the different switches is critical and requires either passive balancing elements, which generate additional losses, or a highly dynamic gate drive and measurement setup for actively balancing the voltages. Furthermore, usually a derating of the switches is necessary due to dynamic overvoltages. In case of a failure, as for example the turn-on of one switch in the series connection is delayed, the resulting overvoltage across the delayed switch could lead to a destruction of the modulator.

On the other hand, connecting several switches in parallel requires a current balancing, which could be achieved, for example, by scheduling the gate signals with an active gate drive as presented in [4]. In case of a failure, one switch would have to conduct a higher current than in nominal case, what, however, is less critical than an overvoltage.

Based on this considerations, a parallel connection of four 1.7kV, 3.6kA IGBT modules and a standard transformer with a turns ratio of 1:170 is needed, as shown in Fig. 4. Since the leakage inductance of the transformer, which limits the pulse rise-time

$$T_{Rise} \sim \sqrt{L_P C_P}, \qquad (1)$$

depends quadratically on the number of turns, the best choice for the number of primary turns is N_P=1. In Eq. 1 L_P denotes the parasitic inductance and C_P the parasitic capacitance of the modulator.

The number of secondary turns is N_S=170 resulting in a relatively large leakage inductance. Furthermore, an additional control circuitry is required in order to obtain/achieve a symmetric current distribution between the parallel connected switches.

By replacing the single standard transformer by a series connection of four transformers with a turns ratio 1:43 and a slightly reduced input voltage of V_{DC}=0.988kV, as shown in Fig. 5a), the current balancing problem could be avoided. There, the secondary windings are all connected in series, so that the secondary current is the same for all four transformers. Due to the Magnetic Flux Law, also the primary currents consequently must be the same, so that the currents are always balanced.

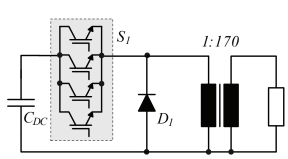

Fig. 4: Parallel connected IGBTs and transformer with a turns ratio of 1:170.

Also the voltage balancing on the secondary and primary is inherently guaranteed, since on the one hand the primary voltages are directly determined by the parallel connected storage capacitors, which are charged up by the same supply, and on the other hand the secondary voltages are given by the turns ratio.

Even in case of failure, if for example one of the four switches is not or delayed turning on, this transformer configuration avoids overloads/overcurrents of single switches. In such a case the output voltage and consequently also the output power is limited to 3/4 of its nominal value in the considered example. The current flowing in the secondary winding of the transformer, whose switch has not turned on, induces a current in the primary circuit, which flows through the freewheeling diode on the primary side, i.e. in the same direction as the switch would be turned on. Therefore, the primary voltage is approximately zero and no power is transfer to the secondary.

Consequently, the parallel/series connection of four individual transformers solves the balancing problem. However, compared to the standard 1:170 transformer a reduction of the parasitic elements respectively of the achievable rise time (cf. Eq. 1) can not be achieved.

The total parasitic inductance of the four transformers is a quarter of the standard transformer's inductance, but the parasitic capacitance is four times higher, so that in total the achievable rise time does not change. However, the characteristic impedance of the transformer ($=\sqrt{L_P/C_P}$) is reduced, which results in a larger overshoot after the rising edge.

The parasitic inductance and capacitance of the transformer are largely determined by the distance between the primary and secondary winding, whereas the distance is given by the maximum electric field respectively the voltage between the primary and secondary.Therefore, the voltage distribution of the four individual transformers is basically the same as the one of the standard transformer, if a parallel connection of all four primaries is assumed.

Considering an Inductive Adder topology [6], where each primary circuit is related to an independent electric potential, the distances between each primary and secondary winding could be reduced, which results in a smaller leakage inductance. There, it is important to minimise the coupling capacitance between all primaries, since the additional coupling capacitance also has to be charged during the rising edge of the pulse.

However, due to the isolation of the primary circuits and related charging power supplies, the circuit complexity significantly increases.

Another method to reduce the leakage inductance is enclosing all four cores by a single secondary winding, so that the flux

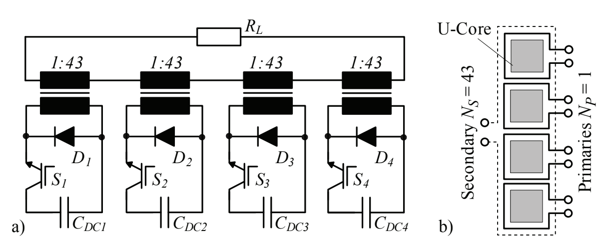

Fig. 5: a) Series connection of 4 transformers with a turns ratio of 1:43 and parallel connected storage capacitors. b) Schematic sketch of a Matrix Transformer with four primaries wound on four separate U-cores and a secondary winding enclosing all four cores.

in the four cores adds up in the secondary winding (cf. Fig. 5b)). Due to adding the fluxes also the voltage induced in the secondary multiplies by four ($V \sim N\mathrm{d}\Phi/\mathrm{d}t$). Therefore, the number of turns on the secondary can be reduced to $N_S = 43$ for generating an output voltage of 170kV. As the secondary winding encloses all four cores, some volume between the primary and secondary winding is saved, as shown in Fig. 6. The saved volume directly results in a reduced leakage inductance compared to the series connection of the standard transformers.

In general, this transformer configuration is called Matrix Transformer [7], where the conversion ratio between the primary and secondary voltage is not only defined by the turns ratio but also by the ratio of enclosed core areas, i.e. enclosed flux shares [7]. There, the voltage ratio is given by

$$\frac{V_S}{V_P} = \frac{N_S}{N_P}\frac{A_S}{A_P}, \tag{2}$$

where V_ν denotes, the winding voltage, N_ν the number of turns and A_ν the area enclosed by winding ν.

Since the primary windings are electrically isolated and therefore only a magnetic coupling via the secondary exists, the DC input voltage could be supplied by a single source again simplifying the circuit design of the modulator system. Besides the Matrix Transformer in literature also Split Core, Fractional Turn [8] transformer as well as Inductive Adders [6] have been proposed. However, these are just special cases of the general Matrix Transformer concept.

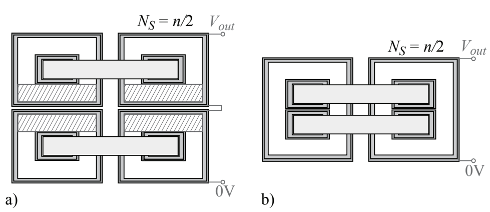

Fig. 6: Saved volume between primary and secondary winding resulting in a reduced leakage inductance.

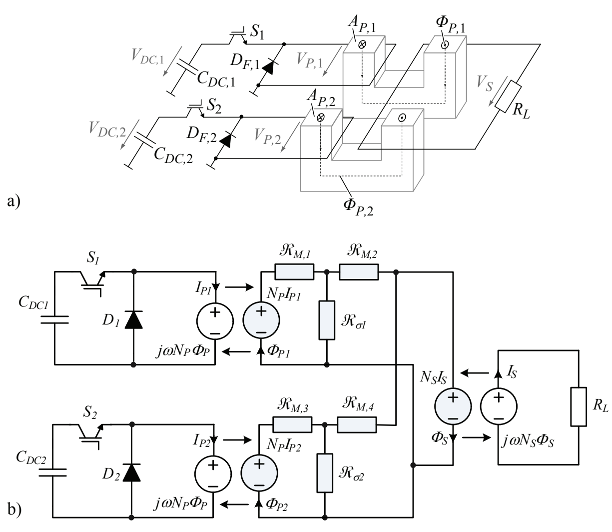

Fig. 7: a) Simplified Matrix Transformer consisting of 2 cores and b) equivalent circuit for the Matrix Transformer based on a reluctance circuit (blue shaded) for the transformer.

Similar to the configuration with four individual transformers, also the special Matrix Transformer with only one winding on each core, has the property of inherent current sharing between the primary circuits ($C_{DC\nu}, S_\nu$ and D_ν), even if the switches do not turn on at the same time, or the parasitic resistances in the circuits are different.

This feature is explained with an equivalent circuit of the Matrix Transformer consisting just of two cores, as shown in Fig. 7a). In the equivalent circuit the transformer is replaced by its reluctance model based on the simplified Ampere's and Faraday's law [9] (cf. Fig. 7b). The quantity $\Re$ represents the magnetic reluctance and the product $N \cdot I$ is known as the magneto motive force (MMF). Based on Ohm's law

$$N \cdot I = \Re \cdot \Phi \qquad (\rightarrow V = R \cdot I)\,, \tag{3}$$

as well as Kirchhoff's voltage (VL) and current law (CL)

$$\begin{aligned} &\text{VL: } N\cdot I = \Phi\cdot(\Re_1+\Re_2\cdots+\Re_n) \quad \text{(for a given path)} \\ &\text{CL: } \Phi_1+\Phi_2+\cdots+\Phi_n = 0 \qquad \text{(for a given node)} \end{aligned} \tag{4}$$

for the magnetic circuit it could be seen, that both primary MMF-sources (N_PI_{P1} and N_PI_{P2}), which correspond to voltage sources in an electric circuit, have to be in the same order of magnitude as the secondary MMF-source N_SI_S in order to limit the flux in the cores. Since for both primary windings the same number of turns is assumed, this results in similar values for the currents, i.e. $I_{P1} \approx I_{P2}$. If one of the primary MMF sources is significantly different compared to the secondary source, the flux in the related core will change rapidly. Due to the changing flux a voltage is induced in the primary winding ($v = Nd\Phi/dt$), which tries to cause a current flow, that balances the MMFs.

In case, where one of the primary switches, S_1 or S_2, turns on later than the other one, the current, which is induced due to the balancing of the MMF sources, flows via the respective freewheeling diode $D_{F,\nu}$. As soon as the delayed switch turns on, the conducting freewheeling diode is hard commutated. Depending on the recovery time of the diode, this could lead to a large reverse recovery current and losses.

In order to avoid this commutation in case of a delayed turn on, each primary circuit could have a winding around each core, as shown in Fig. 8 [10]. However, in case of mismatched turns-on and turn-off times of the switches S_1 and S_2, the switch current can reach twice the nominal current, due to the additional winding. Therefore, with this solution on the one hand the complexity of the transformer design increases significantly and on the other hand an overload of the switch turning on to early could happen. Furthermore, the additional windings are not really necessary, since the switching operation could be synchronised relatively simply by an active gate control, as explained in [4]. There, the rising and falling edges of the switch currents

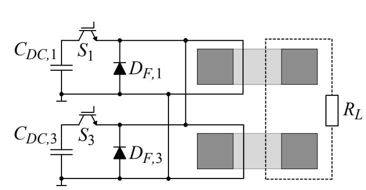

Fig. 8: Cross coupling of primary windings for avoiding turn-on of a freewheeling diode in case the switches do not turn on synchronously.

and voltages are measured and gate signals are scheduled, so that all switches are turned on at the same time. The required current measurement is also used for overcurrent protection of the modulator.

In case of asynchronous switching times, the output voltage of the Matrix Transformer without cross coupled windings is smaller than the nominal voltage during the conduction interval of the freewheeling diode, since the primary circuit with the conducting diode does not generate a flux in the secondary. This could be used to shape the pulse or to compensate the voltage drop of the input capacitors by sequentially turning on the switches in case the modulator has several switches.

Summing up, the Matrix Transformer basically shows the same behaviour as the series connection of standard transformers, shown in Fig. 5a). However, due to the reduced leakage inductance a better pulse performance is achieved.

IV. Reluctance Model

Based on the considerations about Matrix Transformers in Section III, now the transient behaviour of the Matrix Transformer applied in the considered power modulator (cf. Fig. 1 and Table I) is discussed more in detail. In Fig. 9 a photo of the transformer and a circuit diagram of the modulator is given. The transformer consists of two cores instead of four in order to limit the circuit complexity and the magnetising current. Each core carries two primary windings and the four primaries are distributed to the four legs of the U-cores.

The secondary winding encloses both cores in order to add the fluxes, as described in Fig. 7. Additionally, the leakage inductance and the current density in the winding is reduced by connecting two secondary windings in parallel, each enclosing the two legs of the two cores on one side. With the reduction of the leakage inductance, the distributed capacitance is increased, so that in total the achievable rise time is not changed.

A symmetric current sharing between the two primary windings on the same core is not inherently guaranteed, so that a current balancing method is required [4]. With the current balancing method simultaneous switching times are achieved, so that the freewheeling diodes do not turn on and have not to be commutated hard.

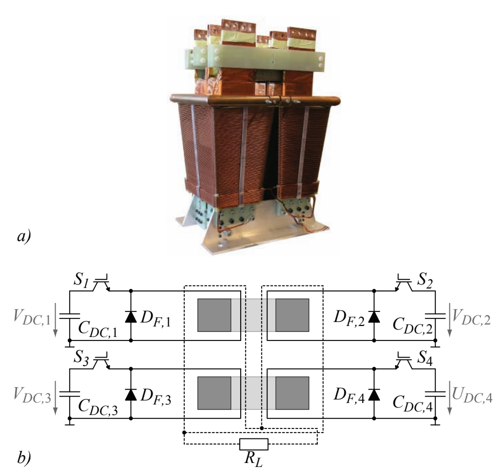

Fig. 9: a) Photo of the transformer and b) schematic of the solid state modulator with two cores, two primary windings around each core and the secondary enclosing both cores.

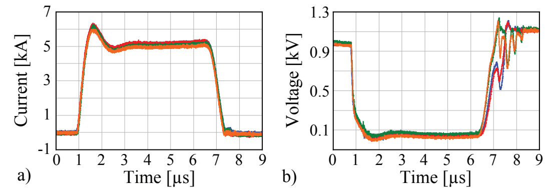

Fig. 10: Measured current a) and voltage b) distribution between the four primary windings of the transformer shown in Fig. 9.

A simplified reluctance model of the Matrix Transformer is shown in Fig. 11. This consists of four MMF sources for the primary windings, one MMF source for the secondary, the reluctances for the core R_{mag}, and the leakage paths $R_{\sigma,\nu}$. The geometric arrangement of the windings is shown by grey, dashed lines for clarifying the model.

The flux distribution for the ideal case, when all four switches turn on simultaneously, is shown in Fig. 12a). There, the two fluxes of the primary winding are added in the secondary winding and the flux in the leakage paths is determined by the MMFs sources of the primary windings.

In the following, the situation of asynchronous switching and the influence of the Matrix Transformer on the voltage/current distribution are considered. Assuming for example, that switch S_1 turns on before the other three switches. In this case, MMF source 1, i.e. $N_{P,1}I_{P,1}$, causes a flux distribution as given in Fig. 12b) until the other three switches also turn on.

Due to the Matrix Transformer configuration $Core$ 2 is magnetically coupled to the primary winding $W_{P,1}$ via the secondary winding, which induces a flux in $Core$ 2, so that the effective magnetizing inductance is doubled, i.e. the two cores are connected in parallel and the effective reluctance seen by $W_{P,1}$ is halved in case nothing would be connected to windings $W_{P,3}$ and $W_{P,4}$. However, in the considered case, the freewheeling diodes are connected across the, what influences the current and

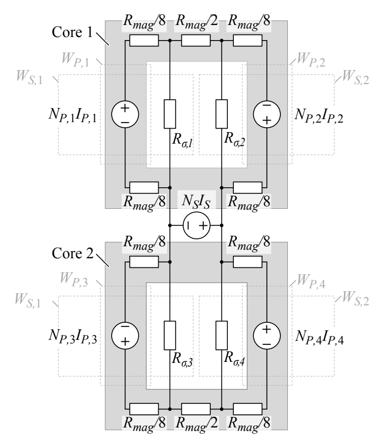

Fig. 11: Reluctance model of the transformer shown in Fig. 9. $W_{P,1}$ to $W_{P,4}$ are the primary windings, $W_{S,1}$ and $W_{S,2}$ the parallel connected secondary windings. $R_{mag} = N_P^2/L_{mag}$ is the reluctance of the magnetic core and R_σ describes the leakage between the primary and the secondary windings. In the model the magnetic coupling between the primary windings on different cores is neglected as it has no influence on the transient behaviour.

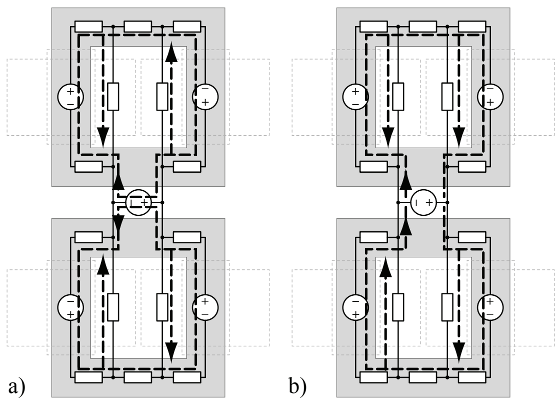

Fig. 12: Flux distribution in the reluctance model (Fig. 11 for a) synchronous turn on of all switches and b) in case switch S_1 is turned on and the three switches are delayed.

the flux distribution. Before this is explained, first the situation for the second winding on $Core$ 1 is considered.

In $Core$ 1 the flux through winding $W_{P,2}$ is in the same direction as it was in the synchronous case, i.e. the induced voltage $W_{P,2}$ has the same polarity as in normal operation and the voltage across switch S_2 decreases. Consequently, the switching losses are reduced due to the reduced voltage as soon as switch S_2 turns on.

The rising edge of the voltage across $W_{P,2}$ is approximately synchronous to the one of $W_{P,1}$, although the rising current edge is delayed until S_2 closes, since there is no alternative current path. The flux in the leakage path $R_{\sigma,2}$, however, is in inverse direction compared to normal operation. Therefore, the current rise in primary circuit 2 is slower, since the flux in the leakage inductance must be reversed.

In windings $W_{P,3}$ and $W_{P,4}$ the situation is different. There, the flux in the windings is in opposite direction and the flux in the leakage path in the same direction compared to the normal operation given in Fig. 12a). The inverse flux in the windings results in an inverted voltage at the winding terminals, which is clamped by the freewheeling diodes. The freewheeling diodes keep the voltage constant and the rate of rise of the flux in $W_{P,3}$ and $W_{P,4}$ to very small values. Winding $W_{P,1}$ sees the two freewheeling diodes D_3 and D_4 parallel connected in series to the load R_L and the effective magnetising inductance seen by $N_{P,1}I_{P,1}$ is equal to the one of $Core$ 1. Due to the freewheeling diodes the current edges are synchronous with S_1 but the voltage edges are delayed until the currents are commutated by the switches S_3 and S_4.

It is important to note, that the resulting current in $W_{P,3}$ and $W_{P,4}$ is in the same direction as in case switches S_3 and S_4 would be turned on, so that the MMF sources have the correct polarity. This and the flux driven by $W_{P,1}$ lead to a flux direction in the leakage paths $R_{\sigma,3}$ and $R_{\sigma,4}$ of $W_{P,3}$ and $W_{P,4}$ which is the same as in the synchronous case, so there is no additional delay in the currents for reversing the leakage flux as with $N_{P,1}I_{P,1}$.

Considering the described situation, it could be seen that the current edges as well as the voltage edges are influenced by the Matrix Transformer and by the turn-on sequence of the single switches. Assuming for example, that S_1 and S_2 turn on at the same time, then all the rising current edges would be synchronous, although S_3 and S_4 have not turned on yet. The rising edges of the winding voltages, however, are not synchronous because not all switches are turned on at the same time. For synchronising the switching transients of all four IGBTs it is therefore necessary to synchronise the current as well as the voltage rising edges, for example by the scheduling concept presented in [4].

Note: In the case described above, the freewheeling diodes started to conduct/clamp the winding, since both switches of the primary circuits mounted on the same core – $Core$ 2 in this case – are delayed. If one of the two switches – S_3 or S_4 in the considered case – turns on at the same time as S_1, the freewheeling diodes would not conduct. Therefore, by having more than one primary circuit connected to a core, the probability that a freewheeling diode has to commutated hard decreases significantly.

V. Conclusion

In this paper, the transient flux distribution in the cores and the voltages across the primary windings during turn-on and turn-off of modulators with Matrix Transformers are investigated based on reluctance models. Also the inherent current edge synchronisation in windings on different cores of the Matrix Transformer is explained in detail. Furthermore, the influence of parasitic capacitances on the overvoltage and switching losses at turn off are discussed.

With the deeper understanding of the Matrix Transformer and its influence on the transients, a better design and a higher performance of solid state modulators with Matrix Transformer are possible.

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