Review of very high frequency power converters and related technologies

Wang, Yijie; Lucia, Oscar; Zhang, Zhe; Guan, Yueshi; Xu, Dianguo

Published in:
IET Power Electronics

Link to article, DOI:
10.1049/iet-pel.2019.1301

Publication date:
2020

Document Version
Peer reviewed version

Citation (APA):
Review of very high frequency power converters and related technologies

Yijie Wang¹, Oscar Lucia², Zhe Zhang³, Yueshi Guan¹, Dianguo Xu¹

¹School of Electrical Engineering & Automation, Harbin Institute of Technology, No. 92, West Dazhi Street, Nangang District, Harbin, People’s Republic of China
²Electronic Engineering and Communications Department, I3A, Universidad de Zaragoza, 50009 Zaragoza, Spain
³Department of Electrical Engineering, Technical University of Denmark, 2800 Kgs. Lyngby, Denmark

Abstract: With the increasing demand for volume reduction and efficiency improvement, very high frequency (VHF) power converters (30–300 MHz) have attracted great interest. Under such high operating frequency conditions, the value and volume of passive components can be greatly reduced, and the power density can be improved. However, many concerns and challenges accompany the increasing operating frequency, such as high switching loss, high magnetic components loss and high driving circuit loss. Including various topologies of the VHF converter, this study reviews the state-of-the-art technology involved in the VHF power converter, also encompassing the inverter stage and matching network stage. Secondly, different magnetic components and semiconductor devices were evaluated under the VHF operating condition. Thirdly, the high-efficiency driving methods, such as the resonant driving method and self-resonant driving method, were demonstrated. A guideline for converter design and system optimisation of the VHF converter and related technologies, including all components and systems, is illustrated in this study. Finally, the future research hot spots and challenges have been pointed out as guidance for further advanced VHF power conversion techniques.

1 Introduction

In many power converter applications, great demands have been put forwards for small volume, easy manufacturability and better performance. To address these concerns, a fundamental method is to increase the operating frequency. Under high operating frequency conditions, the energy stored in passive components during every period of operation can be greatly reduced. Thus, the value and volume of passive components decrease, leading to a high power density based on miniaturisation and integration. Very high frequency (VHF) technique has begun to be adopted in a wide range of applications, including the LED driving system, VRM and wireless power transmission. However, there are many challenges that need to be solved in tens and hundreds of MHz, such as high switching loss, effect of parasitic component, high magnetic loss, and high driving loss, etc. In this review, a comprehensive evaluation of the state-of-the-art VHF technique is presented. A deep research of how these aforementioned challenges have been solved is investigated from the perspective of topology, component and driving strategy. For VHF converters, it can be seen as a merged field between low-frequency power electronics converter and low power radio-frequency (RF) circuit. RF amplifiers are the typical RF circuit, dealing with small DC-to-AC power transformation. Based on the satisfactory soft-switching performance, some amplifier architectures have been applied in VHF converters, such as Class D/E/F [1–8]. In Switching Mode Power Supplies (SMPSS) VHF converters, besides DC-to-AC transformation, DC-to-DC energy transformation is widely needed. Thus, a rectifier stage performing AC-to-DC needs to be investigated. Due to the duality principle, the rectifier stage can be analysed and designed accordingly. Combining various inverter stages and rectifier stages, different sophisticated power converters and their associated control strategies have been proposed, which have been detailed analysed in this paper [9–16]. Moreover, the characteristics and design principles of such topologies have been discussed. Besides proper topologies, the VHF converters put forwards a great demand for semiconductor devices and passive components. It is noteworthy that in the VHF condition, the needed passive resonant components have a quite small value, thus, the parasitic capacitive and inductive components of semiconductor devices and their layout must be carefully considered. On the one hand, the parasitic components are expected as small as possible, making the system absent of their influence; On the other hand, the parasitic components may be taken as the resonant components in power circuits. Thus, a deep evaluation of semiconductor devices is performed in this paper [17–24]. Another challenge for the VHF converter is magnetic components, which lead to a large part of system volume. There are two choices for high-frequency inductors or transformers, i.e. with a magnetic core and without magnetic core (air core). With the help of the magnetic core, the volume can be greatly reduced compared with the air core one under the same inductance in ideal condition, as well as the magnetic field can be well controlled. However, the loss and thermal performance must be taken into consideration. Thus, this paper provides a deep comparison as guidance for magnetic component selection under VHF conditions [25–31]. Besides the perspectives of topology and component selection, in VHF conditions, the system efficiency is also significantly affected by driving loss. In the square-waveform driving method, the switch turns on and turns off by charging and discharging input capacitors during every cycle. However, the energy of the switch input capacitor is totally dissipated. Thus, the driving loss goes to a relatively high level with the increment of operating frequency, especially when it reaches tens of MHz. To solve this problem, advanced driving methods, such as the resonant driving method and the self-resonant driving method are proposed, which are analysed in this review as well [32–37]. In this paper, VHF approaches are analysed from several perspectives, such as topology, component and driving method etc. In Section 2, different inverter stages, rectifier stages and matching networks are studied. Also, the non-isolated and isolated converters are introduced. VHF magnetic components and semiconductor device characteristics are presented in Section 3. In Section 4, advanced high-efficiency driving methods are demonstrated in VHF conditions. Section 5 discusses the future research hot spots.
and challenges and finally, Section 6 draws the conclusions of this paper.

2 Topology analysis of VHF converter

2.1 Non-isolated DC/DC topology

Resonant power converters with soft-switching characteristics are preferred in VHF operation. According to current research lines, three topologies are mainly adopted in the VHF applications, namely the class DE [38–40], SEPIC [41–43], and class E converters [44, 45], as shown in Fig. 1, respectively. These topologies can be divided into inverter stages and rectifier stages, which can achieve DC-to-AC transformation and AC-to-DC transformation, respectively. The inverter and rectifier stages based on the class DE and class E can be assembled according to different situations. For the class DE converter, there is only one resonant inductor adopted in the circuit, which helps to reduce the system volume. However, it can be seen that there are one low-side switch and one high-side switch in the inverter stage. For manufacturing, some advanced processes, such as triple well or silicon-on-insulator (SOI), should be adopted. Meanwhile, great attention should be paid to the parasitic components in the half-bridge structure, which significantly affects the operation of the VHF converters.

As Fig. 1b shown, only one low-side ground-referenced switch is needed in the SEPIC converter. For the input side, it can be seen as a class E inverter, and the main waveforms are shown in Fig. 2. It can also be seen that there is one high side diode in the rectifier stage. Schottky diodes with Si and SiC materials are widely used in VHF converters due to the low forward voltage drop and fast switching speeds. However, in VHF situations, the forward recovery voltage significantly affects the characteristics of converters. In a very short transition time, the forward voltage can be increased by 50%, causing unexpected loss and reducing system efficiency [46]. In high output voltage, the forward voltage drop plays a small role; however, in low output voltage condition, the conduction loss cannot be ignored. Furthermore, for VHF of the order of hundreds of MHz, diodes with CMOS design methods are rarely available. At very high frequencies, the conductivity modulation of power diodes causes an inconvenient loss, which has been analysed in [47, 48]. Thus, the diode in the rectifier stage is expected to be located at the low side, which can be replaced by a switch with a simpler driving circuit.

The class E converter as shown in Fig. 1c is the most widely adopted topology among all DC–DC VHF converters. It can be seen that with one low-side switch in the inverter stage and one low-side diode in the rectifier stage, the topology is very suitable for operation in the VHF condition and can be easily integrated. As mentioned above, in the rectifier stage, a synchronous transistor can be adopted to replace the diode. In addition, the output parasitic capacitance of the switch and diode can be absorbed by the corresponding resonant capacitors. Even under higher frequency conditions, the output capacitance is sufficiently large and no more discrete resonant capacitors are needed. However, it is remarkable that the values of output capacitance of switches and diodes are not constant, which usually reduces with increased drain-to-source voltage and forms a non-linear relationship over voltage. Thus, the non-linear characteristics should be considered in the design procedure. Moreover, the capacitance across the switch need to be selected beforehand, and cannot be larger than the parasitic value. Based on the capacitance, the corresponding resonant inductor can be decided upon, and hereby to determine the operating frequency. The switch of class-E converters needs to lack more than twice the input voltage, therefore to reduce the switch voltage stress further. Class $\Phi_2$ circuit is proposed as shown in Fig. 1d, the main waveforms are shown in Fig. 3. It can be seen that the switch voltage can be greatly reduced with the help of the third harmonic. As the circuit shows, an $L$–$C$ branch is added in parallel with the switch drain–source. This branch can be used to adjust the harmonic of the voltage waveform. Table 1 shows the characteristics comparison among these converters.

The matching networks are added between the inverter stages and rectifier stages to adjust the equivalent impedance of the...
rectifier stage. Fig. 4 shows the typical matching networks consisting of inductors and capacitors. Based on the frequency domain characteristics, they can be divided into the low-pass type and high-pass type. The low-pass type can eliminate the effect of high harmonics. However, from the perspective of energy adoption, besides fundamental waveforms, the high-pass type can also take advantage of high harmonics as well. Thus, it helps to improve system efficiency. The L-type matching network comprises a capacitor and an inductor, as shown in Fig. 4. The relationship between $Z_L$ and $Z_R$ can be regulated using a matching network. In addition, the voltages of the input and output port can be changed. Thus, the aforementioned matching network can be seen as a non-isolated transformer. At the nominal operating point, the impedance $Z_L$ and $Z_R$ are both resistive. However, a limitation of the L-type structure is that with the variation of $Z_R$, the impedance of $Z_L$ comes into inductive or capacitive, as shown in Fig. 5. It is not conducive to keep soft-switching characteristics in the inverter stages.

To address this problem, a T-type matching network is proposed, which is shown in Fig. 6. It is composed of one inductor and two capacitors. With the proposed structure and optimal parameter design, the impedance of $Z_L$ can retain resistive characteristics regardless of the change in $Z_R$. Therefore, it is easier to maintain the operation of the inverter switch operates in soft-switching conditions. Fig. 7 shows the impedance angle curves with the variation of $Z_R$. It is seen that, if the capacitors are chosen to have the same value, the impedance angle can be retained at zero, even with a change in $Z_R$. According to the duality principle, the π-type network can also be proposed as shown in Fig. 8. As shown in Fig. 9, the same characteristics can be achieved as previous results. With the addition of one capacitor, the two matching networks can achieve high efficiency over a wide range of load variations. However, though the T-type or π-type can keep

<table>
<thead>
<tr>
<th>Topology</th>
<th>Switch number</th>
<th>Diode number</th>
<th>Inductor number</th>
<th>Capacitor number</th>
<th>Voltage stress ($V_d/V_{in}$)</th>
<th>Driving complexities</th>
</tr>
</thead>
<tbody>
<tr>
<td>SEPIC</td>
<td>1</td>
<td>1</td>
<td>2</td>
<td>2</td>
<td>1</td>
<td>difficult</td>
</tr>
<tr>
<td>class E</td>
<td>1</td>
<td>1</td>
<td>3</td>
<td>3</td>
<td>3.5</td>
<td>easy</td>
</tr>
<tr>
<td>class $\Phi_2$</td>
<td>1</td>
<td>1</td>
<td>4</td>
<td>4</td>
<td>2.6</td>
<td>easy</td>
</tr>
</tbody>
</table>
resistive transformation characteristics, however, the transformation ratio varies under different ZR conditions.

To solve the aforementioned problem, a high-order matching network is also proposed. As shown in Fig. 10, two high-low bandpass matching networks are presented. Based on the optimal design method reported in [49], the high-low pass matching network can maintain resistive transformation characteristics at the desired resonant frequency even if the equivalent load is increased. Moreover, at the resonant frequency, the voltage transformation ratio can retain a constant value. Thus, it can be understood that for the VHF converters with the same inverter and rectifier circuits, the proposed high-low matching network can be applied to achieve a synchronous structure with the same driving signal. As shown in Fig. 11, the bidirectional and synchronous VHF converter based on the class \( \Phi_2 \) inverter and rectifier stage is proposed.

One shortcoming for the aforementioned non-isolated matching networks is that they are very sensitive to the operating frequency. A small frequency deviation will lead to transformation characteristics variation. Also, the accuracy of passive components is expected to be at a high level, which helps to achieve expected transformation characteristics. Thus, in future work, the high-performance non-isolated matching network should be further investigated. On the other hand, the inverter stage, which can guarantee soft-switching characteristics within a wide load range, should be studied.

2.2 Isolated DC/DC converter topology

Besides the non-isolation structure, the isolation function is expected in many application fields. As shown in Fig. 12, a VHF converter based on the class \( \Phi_2 \) inverter stage is achieved with the capacitive isolation barrier method. With an additional capacitor in the return loop of the rectifier stage, the isolation function can be effectively achieved. Under operating frequencies of hundreds of kHz, the value and volume of the isolation capacitor must be very large. However, in VHF conditions with reduced energy requirements, the value and volume can be significantly decreased. Also, the parasitic resistance and volume of the ceramic capacitor are usually quite lower than its magnetic component counterpart, which helps to improve the system efficiency and power density. Thus, for low voltage isolation requirements, the VHF converters with the capacitive isolation method can be applied.

However, for higher voltage isolation requirements, significant problems exist in the capacitive isolation method. One problem is that with the increment of the rated voltage, the volume of the capacitor greatly increases, which is not conducive to reduce the converter volume. Furthermore, with a large capacitor size, the parasitic inductance caused by leads also greatly affects the VHF operation. It should be mentioned that in Fig. 12, the parasitic inductance of CRES can be absorbed by LRES; however, the parasitic inductance of CRES1 leads to great impact. Another problem is that for KV isolation applications, there are almost no high-quality-factor capacitors that are available to operate at such high switching frequency. With a low-quality factor, the system suffers significant losses.

In addition, for the circuit shown in Fig. 12, it can be seen that an additional capacitor is needed. For power density sensitive or cost-sensitive applications, the component number is expected to be as small as possible. Meanwhile, in very high-frequency applications, increasing the number of components leads to more tracks and more component leads, which causes unexpected parasitic inductances. Thus, the VHF operation appreciates a topology with a small number of components.

Similar to the low-frequency conditions, magnetic isolation based on the transformer is the most common method. It can be applied to both the low voltage and high voltage applications. Besides transformer-based solutions, some problems need to be addressed. One problem is that the leakage inductance and magnetising inductance affect the operating mode of VHF converters. At low operating frequency, a transformer is usually expected to be as ideal as possible by minimising leakage inductance and maximising magnetising inductance. Based on magnetic cores with high permeability, large magnetising inductance and small magnetising current can be achieved. Meanwhile, with advanced winding methods, a very high coupling coefficient can be realised in low operating frequency. Also, the resonant inductor value is quite large in hundreds of kHz. Given that, the transformer can be taken as an ideal one, which only transfers energy from the primary side to the secondary side.

However, under VHF conditions, magnetic cores lead to an unacceptably large loss. With the increase of operating frequency, the air-core transformer with small magnetising inductance is gradually applied in VHF conditions. Without any magnetic core, an air-core transformer has loosely coupled coefficients. In addition, in VHF converters, the necessary resonant inductor values are in the order of tens or hundreds of nH so that the influence caused by leakage inductance and magnetising inductance cannot be ignored any more. One effective way is to consider the leakage inductance and magnetising inductance as resonant components through the optimal topology design.

Fig. 13 shows an example of optimising the topology of an isolated VHF converter. An isolated DC/DC converter based on the class \( \Phi_2 \) inverter stage and class E rectifier stage is illustrated in Fig. 13a, where one transformer is added between these two stages. The leakage inductance in the secondary side can be combined with the resonant inductor. From the perspective of AC analysis, the DC input source can be seen as a short circuit, thus, one port of the transformer primary side can be adjusted to the input side, as shown in Fig. 13b. Within the dashed-line box, it can be seen that the resonant inductor LF can be combined with the transformer primary side leakage inductance. Finally, as shown in Fig. 13c, the magnetising inductance merged with input inductance. Therefore, this example shows the approach of adopting the leakage inductance and magnetising inductance into the VHF isolated converter. It should be mentioned that in Fig. 13a, capacitor CRES plays the role of blocking DC voltage. Thus, its value is much larger than the resonant capacitor. The voltage across CRES can be seen as constant and therefore, this capacitor can behave as a voltage source. In Fig. 13b, there is no need for DC voltage blocking and no CRES exists. Compared with Figs. 13a and c, two resonant inductors are saved, which can help to reduce the count and volume of the components, in order to improve the power density. In addition, the inductors loss can also be avoided.

Based on the aforementioned converter, the diode in the secondary side can be replaced by a synchronous switch as shown.
optimal winding structure with reduced transformer inter or intra parasitic capacitance should be investigated.

Table 2 shows some typical VHF converters performance, including power rating, input/output voltage, efficiency and power density. The power densities of most VHF converters can be around 100 W/inch$^3$. The VHF converters mainly focus on the low power applications with tens of volt input and output voltage and tens of watt power. The trend of the VHF converter is to further improve the power and voltage levels. A comparison with modern non-VHF converters is also shown in Table 3, including numerical comparative values of power rating and efficiency. It can be seen that there is a gap for VHF converters to be applied in higher power situations.

3 Magnetic and switch components

3.1 VHF magnetic component with core

In a VHF converter, the magnetic component is an important part to deal with. In general, the value of inductors and transformers forms an inverse proportional relationship with the operating frequency. However, we cannot conclude that the components volume reduces continuously with an increase in operating frequency. The volume scaling is also related with winding loss [57–59], core loss and permeability [13, 15–18, 60–62], and heat transfer [63, 64] under various operating conditions. Research on the relationship between size and frequency has gained increased attention. In [60], under a certain operating frequency, the inductor quality factor with different kinds of loss is deeply investigated. In [13] with limited heat transfer capability, the volume characteristics have been analysed with various frequencies. Furthermore, in [64], with more efficiency limitations, the property of transformer size has been researched. In [22], the transformer design method considering the core loss and winding loss is proposed under different operating frequencies. Besides volume scaling characteristics, the power density property has been investigated, considering the above said loss and dissipation characteristics. By
research, it is demonstrated that because of the loss and heat limitations, the volume cannot constantly reduce, even if the operating frequency continuously increases.

For magnetic components with cores, the core loss is significant for volume minimisation. For VHF conditions, coreless magnetic components can be utilised, which suffers only from winding loss. Thus, a balance between core loss and winding loss is required, when choosing the magnetic component type.

Under a certain inductance and loss situation, the quality factor of an inductor can be obtained as follows [65]. Here, $\varepsilon$ represents the linear scaling factor.

$$ Q = \frac{2\pi f L}{R_{ac}} = \frac{2\pi f N^2 K \varepsilon}{N^2 K \varepsilon \sqrt{f}} = \frac{K \varepsilon}{\sqrt{f}} $$

where $L$ represents the inductance, $N$ represents the turns number, $R_{ac}$ represents the inductor $AC$ resistance, $\varepsilon$ is the linear dimension factor, and assuming $L = K \varepsilon$, $R_{ac} = K \varepsilon \sqrt{f}$.

It can be seen that the quality factor forms a proportional relationship with $\varepsilon$ and frequency $f$. Therefore, to maintain a constant inductance and quality factor at various frequencies, the linear dimension can be scaled as $f^{-1/2}$, thus, the total volume varies by $f^{-3/2}$.

Without considering loss limitation, the inductor volume of a certain value constantly decreases at the rate of $f^{-1/2}$ when the operating frequency increases. However, in fact, for a real application, there must be core loss or winding loss. With volume reduction, the heat dissipation ability weakened and it is easy to reach the thermal and operating temperature limitations.

### 3.2 VHF air-core magnetic component

To reduce core loss, the air-core magnetic component has been widely adopted in VHF conditions. There are different winding structures for air-core magnetic components, such as spiral, solenoid and toroidal. Compared with cored components, air-core components usually need more turn numbers to achieve the same inductance.

As discussed in [65], Fig. 17 shows different volume curves with various inductor structure and core materials, such as high permeability magnetic material 3F3, low permeability RF material P, and air-core structure. A small difference among these three conditions is that the air core structure is unshielded, while the two cored structures are magnetically shielded. From the curves, it can be seen that at low frequency, when the inductor has a significant heat dissipation ability, the volume of the two cored inductor can be scaled as $f^{-1/2}$. However, when the temperature becomes a dominant factor, the volume of the inductor can be scaled as $f^{-3/2}$. It can be seen that the previous analysis can effectively capture the inductor characteristics. Nevertheless, after reaching a certain operating frequency, due to the loss and temperature limitation, the volume of these two inductors with cores increases with the rise of operating frequency. However, for the air core structure, it can be seen that it decreases constantly as operating frequency increases.

### Table 2: Performance of typical VHF converters

<table>
<thead>
<tr>
<th>Converter</th>
<th>Frequency, MHz</th>
<th>Power, W</th>
<th>Input/output voltage</th>
<th>Efficiency, %</th>
<th>Power density, W/inch³</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cai et al. [50]</td>
<td>30</td>
<td>50</td>
<td>36 V/24 V</td>
<td>73.3</td>
<td>100</td>
</tr>
<tr>
<td>Madsen et al. [51]</td>
<td>46</td>
<td>5.7</td>
<td>230Vac/15 V</td>
<td>78</td>
<td>146</td>
</tr>
<tr>
<td>Guan et al. [11]</td>
<td>20</td>
<td>10</td>
<td>12 V/5 V</td>
<td>80</td>
<td>150</td>
</tr>
<tr>
<td>Gu et al. [52]</td>
<td>13.56</td>
<td>200</td>
<td>210 V/30 V</td>
<td>90</td>
<td>10</td>
</tr>
</tbody>
</table>

### Table 3: Performance of modern non-VHF converters

<table>
<thead>
<tr>
<th>Converter</th>
<th>Frequency, MHz</th>
<th>Power, W</th>
<th>Efficiency, %</th>
<th>Power density, W/inch³</th>
</tr>
</thead>
<tbody>
<tr>
<td>Tang et al. [53]</td>
<td>210 kHz</td>
<td>6.6 kW</td>
<td>95.73</td>
<td></td>
</tr>
<tr>
<td>Liu et al. [54]</td>
<td>720 kHz</td>
<td>1.2 kW</td>
<td>96</td>
<td></td>
</tr>
<tr>
<td>Fei et al. [55]</td>
<td>1 MHz</td>
<td>800 W</td>
<td>97.2</td>
<td></td>
</tr>
<tr>
<td>Ahmed et al. [56]</td>
<td>1.6 MHz</td>
<td>250 W</td>
<td>97</td>
<td></td>
</tr>
</tbody>
</table>

![Fig. 17](image1.png) **Fig. 17** Comparison between conventional magnetic material (3F3), RF material (P) and coreless inductor volume [65]

![Fig. 18](image2.png) **Fig. 18** Simplified device model in a lumped form
loss, which forms a proportional relationship with a square parasitic capacitance which varies significantly with the change in these two resistors is affected by the operating frequency. This is analysed and their relationship with operating frequency can be investigated because the corresponding branch impedance is influenced by the soft-switching characteristics of switch. With the lumped switch model, the switching loss and conduction loss can be explained and their relationship with operating frequency can be investigated. Here, the impedance of switch conduction resistance does not change for different frequencies. Thus, the conduction loss is only dependent on the duty cycle and the corresponding current. Thus, the conduction loss is increased under the sinusoidal waveform driving signal. This is because with a resonant driving method, the driving circuit loss plays a dominant role in total system loss. For frequencies of the order of hundreds of kHz, the square-wave driving method is the most widely adopted one. This indicates that a square-waveform signal is used to charge and discharge the switch input capacitance in order to utilise the capacitor-stored energy during every period is completely dissipated. Thus, in very high-frequency conditions, resonant driving methods are gradually adopted.

As shown in Fig. 20, in the driving circuit, a resonant inductor is added in series with the switch gate, which can resonate with the switch input capacitance in order to utilise the capacitor-stored energy. The resonant driving method can greatly reduce losses compared with the square-wave driving method, as shown in Fig. 21. However, it should be mentioned that although the driving circuit loss based on the resonant driving method can be greatly reduced compared with square-wave driving, the switch conduction loss is increased under the sinusoidal waveform driving signal. This is because with a resonant driving method, the driving signal has a sinusoidal waveform, thereby making the rising edge and falling edge slower. Thus, during these times of transition, the switch on-resistance is high, which results in a higher conduction loss. By increasing the driving voltage amplitude, the switch fully

Table 4  Dependence of device loss mechanisms on device parameters and frequency scaling

<table>
<thead>
<tr>
<th>Mechanism</th>
<th>Device dependence</th>
<th>Frequency dependence</th>
</tr>
</thead>
<tbody>
<tr>
<td>conduction loss</td>
<td>α $R_{ds,on}$</td>
<td>independent</td>
</tr>
<tr>
<td>displacement loss</td>
<td>α $R_{oss}C_{oss}$</td>
<td>$\propto f_s^2$</td>
</tr>
<tr>
<td>gating loss</td>
<td>α $R_gC_{iss}$</td>
<td>$\propto f_s^3$</td>
</tr>
</tbody>
</table>

Some Si Schottky diodes have been found to have poor resonant rectifier performance at VHF, showing higher temperature increases and lower current limits. The applicability of Si Schottky diodes for the VHF resonant rectifier is discussed [66]. The performance of each diode is evaluated by measuring the power loss of the diodes operating in the class E rectifier. The method of thermal characterisation is used to simplify the measurement of power loss. Fig. 19 is the block diagram of the experimental setup for the diode performance evaluation. The power loss of the diode at VHF is determined by measuring the temperature rise of the diode under RF operation. In [66], 14 commercially available Si Schottky diodes are tested at 30 MHz and three of the best-performing diodes are also tested at 50 MHz in [66]. The experimental conclusion is that when operating at 50 MHz, the power consumption is slightly higher, but these three diodes are still useful in the rectification at VHF. As the frequency increases, the value of $C_s$ decreases and approaches the value of the diode capacitance $C_{DSS}$. It is important to use diodes with low intrinsic capacitance in order to maintain the resistance characteristics while maintaining the increased operating frequency.

With the fast development of wide bandgap devices, the SiC and GaN transistors and diodes show superior characteristics compared with the conventional Si devices. With higher carrier mobility, and a certain on resistance, the devices with wide bandgap material can be manufactured with a smaller area, which helps to reduce the parasitic capacitance. Thus, in the VHF condition, the loss can be reduced. Owing to outstanding advantages, the RF and power switches based on wide bandgap material have been the research hotspot in recent years. With further advanced investigation and optimal design strategy, the GaN and SiC devices can achieve better characteristics, which can further improve the performance of VHF converters. However, the output capacitance loss shows different performances in VHF operating situations.

4 Driving method for VHF converters

Besides the research of power conversion topologies, power semiconductors and magnetic components in VHF converters, high efficiency driving circuits for VHF converters also have attracted lots of attention. It is well known that the driving circuit loss forms a proportional relationship with switching frequency. Thus, when the system operating frequency increases to the order of tens and hundreds of MHz, the driving circuit loss plays a dominant role in total system loss. For frequencies of the order of hundreds of kHz, the square-wave driving method is the most widely adopted one. This indicates that a square-waveform signal is used to charge and discharge the switch input capacitance in order to turn the switch on and off. However, by utilising this driving method, the gate charged and discharged energy during every period is completely dissipated. Thus, in very high-frequency conditions, resonant driving methods are gradually adopted.

As shown in Fig. 20, in the driving circuit, a resonant inductor is added in series with the switch gate, which can resonate with the switch input capacitance in order to utilise the capacitor-stored energy. The resonant driving method can greatly reduce losses compared with the square-wave driving method, as shown in Fig. 21. However, it should be mentioned that although the driving circuit loss based on the resonant driving method can be greatly reduced compared with square-wave driving, the switch conduction loss is increased under the sinusoidal waveform driving signal. This is because with a resonant driving method, the driving signal has a sinusoidal waveform, thereby making the rising edge and falling edge slower. Thus, during these times of transition, the switch on-resistance is high, which results in a higher conduction loss. By increasing the driving voltage amplitude, the switch fully
The aforementioned resonant driving method is composed of an oscillator and several paralleled inverters. In order to further simplify the necessary components, a self-resonant driving circuit is proposed. Fig. 24 shows a VHF self-resonant driving circuit based on a series resonant inductor [68]. In the circuit, LG is the resonant inductor and \( V_{bias} \) represents the bias DC voltage. Based on the inductor and the switch parasitic capacitors, a high-pass filter with the capacitive load is formed. The transfer function \( V_{ds}/V_{gs} \) needs to be carefully designed to satisfy the requirements.

Fig. 25 shows the waveform diagram of the switch gate and drain voltages. Based on the analysis detailed in the previous power conversion architecture, the switch drain-to-source voltage is usually in a half-sinusoidal form. Thus, it can be seen that when the driving voltage is low, the switch is off and the drain voltage is high. Meanwhile, when the driving voltage is high, the switch is on and the drain voltage is low. Thus, there should be an approximately 180° phase difference must exist between switch gate and drain voltages.

Fig. 26 shows the Bode diagram of the feedback network with different series inductance parameters. As can be seen from the figure, the network can achieve about 180° phase difference within a certain range of frequency. By changing the value of inductances, the voltage gain at the operating frequency can be adjusted to meet the amplitude requirement of different switches. The bias voltage \( V_{bias} \) can be adjusted to change the switch duty cycle with different threshold voltages. Usually, the duty cycle of the switch is designed to be 0.5, which means that the bias voltage should be around the threshold voltage of the switch. Thus, for different switches, the bias voltage and inductor value should be modified.

However, for the aforementioned self-driving method, besides the switch parasitic capacitance parameters, the series inductor is the only variable which can be varied. Thus, the characteristics of the self-driving network are mainly determined by the switch. With different parameters, some switches cannot meet the self-driving requirement even within a wide inductor range. Therefore, to address this problem, an additional LC branch can be added between switch drain-to-gate or gate-to-source. Figs. 27 and 28 show the equivalent self-driving networks with LC branch in different locations. Using the same method, the transferring functions between switch drain and gate voltages can be obtained. It can be seen that many variables are available to modify the characteristics to meet the magnitude and phase requirements of the self-driving circuit in single-switch VHF converters.

Meanwhile, with the LC branch, the current flowing through the driving circuit can be redistributed compared with the basic situation. The LC branch provides an additional loop for driving the self-driving network are mainly determined by the switch. With different parameters, some switches cannot meet the self-driving requirement even within a wide inductor range. Therefore, to address this problem, an additional LC branch can be added between switch drain-to-gate or gate-to-source. Figs. 27 and 28 show the equivalent self-driving networks with LC branch in different locations. Using the same method, the transferring functions between switch drain and gate voltages can be obtained. It can be seen that many variables are available to modify the characteristics to meet the magnitude and phase requirements of the self-driving circuit in single-switch VHF converters.

Fig. 22 Multi-resonant driving circuit

Fig. 23 Simplified driving voltage with the fundamental component and third harmonic component

Fig. 24 Circuit of a self-resonant VHF driving circuit

Fig. 25 Diagram of switch gate and drain voltage

turned-on time can be increased, which helps to reduce the corresponding loss. However, the driving loss increases in turn with a higher driving voltage. There is a trade-off that exists between the driving loss and conduction loss according to the different system operating conditions. In general, the resonant driving method should be adopted in the low current situation, and the square-waveform driving method should be adopted in the large current situation. Also, from a driving loss perspective, the wide bandgap semiconductor devices with small input capacitance and on-resistance perfectly fit in VHF converters.

To solve the problem above and make a sharper rising or falling edge, the high-order harmonics injection approach can be adopted and the multi-resonant network is proposed, which is shown in Fig. 22 [67]. Compared with the circuit in Fig. 20, it can be seen that an additional inductor and capacitor branch LMR and CMR is paralleled with series resonant inductor LF. The LC branch is added to introduce the third harmonic into the driving waveform. The corresponding simplified driving voltage with the fundamental component and the third harmonic component is shown in Fig. 23. It can be seen that using the third harmonic, the driving signal has a trapezoidal-shape waveform, which makes the rising and falling edge steeper. The multi-resonant network resonates with input capacitance \( C_{iss} \) during every period, thus, part of the energy stored in the input capacitance can be recycled.

The corresponding simplified driving voltage with the fundamental component and third harmonic component is shown in Fig. 23. It can be seen that using the third harmonic, the driving signal has a trapezoidal-shape waveform, which makes the rising and falling edge steeper. The multi-resonant network resonates with input capacitance \( C_{iss} \) during every period, thus, part of the energy stored in the input capacitance can be recycled.
above analysis, Table 5 summaries a characteristics comparison of different VHF driving methods.

### Table 5: Comparison of different VHF driving methods

<table>
<thead>
<tr>
<th>Mechanism</th>
<th>Driving loss</th>
<th>Conduction loss</th>
<th>Complexity</th>
<th>Cost</th>
</tr>
</thead>
<tbody>
<tr>
<td>square-wave driving</td>
<td>high</td>
<td>low</td>
<td>medium</td>
<td>medium</td>
</tr>
<tr>
<td>sinusoidal-wave</td>
<td>low</td>
<td>high</td>
<td>high</td>
<td>medium</td>
</tr>
<tr>
<td>multi-resonant</td>
<td>low</td>
<td>low</td>
<td>highest</td>
<td>high</td>
</tr>
<tr>
<td>self-resonant</td>
<td>low</td>
<td>high</td>
<td>lowest</td>
<td>low</td>
</tr>
<tr>
<td>self-resonant with</td>
<td>low</td>
<td>high</td>
<td>low</td>
<td>low</td>
</tr>
<tr>
<td>LC</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

introduce high-order harmonics. Multiple LC branches with different resonant frequencies can be paralleled, making the driving voltage closer to the trapezoidal or a square waveform. The short turn-on and turn-off transient time can reduce switch conduction loss. For half-bridge very high-frequency converters, a similar method can be adopted to drive the switches. One challenge is that the phase of the driving signals should be carefully designed to obtain the complimentary gate signals with deadtime. Based on the
Resonant topologies with lower switch voltage stress are the attractive point that can widen the system input and output voltage range. It is well known that the switch voltage stress of half-bridge structure is the same as input voltage, which is a significant reduction compared with that of the single switch structure. However, there are some challenges for half-bridge VHF converters. One is the high-side driver; the high-performance VHF half-bridge driver IC should be developed. Another is the parasitic components in the bridge middle point. In VHF, the parasitic components especially inductance, leads to huge voltage oscillations in the gate and drain. Thus, the integrated VHF half-bridge switch module is highly expected. Also, the high-resolution VHF driving signal is a great challenge. To achieve soft-switching characteristics, proper deadtime must be guaranteed, which should be only hundreds or thousands picoseconds. Thus, a high-precision VHF control chip should be developed.

Besides advanced topologies, the driving strategy should be further investigated. With the advantages of low input capacitance and low on-resistance, GaN HEMTs begin to be adopted in HF and VHF converters. One characteristic of GaN HEMTs is that the gate driving voltage should be in a very narrow, usually is −2 to +6 V. However, as analysed above, the resonant driving method is widely adopted in VHF converter which may exceed the driving voltage range of GaN HEMTs. Thus, the high efficiency and high-reliability driving strategy for GaN HEMTs should be developed. Also, for the existing synchronous driving method, the driving loss is still at a high level and the driving circuit is in a quite complex situation. Thus, the simple structure high-efficiency synchronous driving method is expected to be proposed. Meanwhile, for isolated VHF converter, the self-driven synchronous driving method should be developed.

To further utilise the parasitic components, research on the estimation and control methods of parasitic inductance can help to avoid using the discrete inductors. However, how to have a fast and smarter way to control or calibrate them in mass production is a difficult issue. New material and structure are aimed at to achieve the smaller size and higher quality factor, which makes the magnetics easy to be integrated. 3D printed inductors and transformers with lighter weight can also strengthen the merits of VHF converters.

To improve the output power of VHF converters, multiphase paralleled structure and the current balance strategy should be investigated. Also, the integration technology of power devices, driving and control circuits should be promoted, the ultimate goal is to integrate total system into a power IC. Also, how to fully apply the high bandwidth or high dynamic performance which provided by VHF operation should be solved by using advanced control unit and method. EMI issues and cost reduction should be further investigated in further work.

6 Conclusion
This paper has provided a detailed analysis of the VHF converter and the corresponding technologies. The first concern of the VHF converters is suitable topologies. Different VHF topologies with good soft-switching characteristics were demonstrated. Different inverter and rectifier stages were adopted to achieve VHF converters for different input, output and other various application fields. In addition to topologies, the magnetic and switch components were introduced. In VHF conditions, the air core or low permeability core structure magnetic components demonstrated good performance, which helps to reduce the system volume. Furthermore, the switches of wide bandgap material are expected in VHF converter to reduce corresponding loss. The parasitic components should be importantly considered and integrated. Finally, the resonant driving and self-driving methods were analysed. These methods can greatly reduce the driving loss. With these technologies, high performance VHF converters find a wide range of applications.

7 References
[9] Zan, X., Avestruz, A.: ‘Activation at 100 MHz for 12 W VHF wireless power transfer’. 20th Workshop on Control and Modeling for Power Electronics (COMPEL), Toronto, ON, Canada, 2019, pp. 1–8