Measurement Techniques of Common Mode Currents, Voltages, and Impedances in a Flyback Converter for Radiated EMI Diagnosis

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Abstract—For the radiated electromagnetic interference pre-compliance diagnosis in power electronics systems, the measurement of common mode (CM) currents, impedances, and voltages are essentially important. In this article, two measurement issues are first identified in CM current and impedance measurement. The first is the coupling between the cables connected to a power converter and the coaxial cable connected to the measurement equipment. It degrades the measurement accuracy. The second is the coupling between the converter’s attached cables and the ac grid. It degrades the repeatability of measurements. To improve the measurement accuracy and repeatability, this article proposes an improved measurement technique to address these two issues. Furthermore, for the measurement of the voltage difference between primary and secondary grounds, this article also identifies two issues. The first is the poor measurement resolution for the low noise voltage in the high-frequency range due to the high magnitude of line-frequency voltages. The second is that the loading effects of the probe can reduce the measurement accuracy. This article proposed a technique to address these two issues. Finally, the proposed measurement techniques were validated on a flyback converter.

Index Terms—Common mode (CM), electromagnetic interference (EMI), flyback converter, power converter, radiated emission.

I. INTRODUCTION

WITH the development of high power density and high switching frequency power converters, and the applications of wide bandgap devices to power conversions, electromagnetic interference (EMI) is becoming an increasingly serious problem. Power converter products need to comply with the international EMI regulation standards, e.g., EN55022 and CISPR 22, before they are allowed to be sold on the market. Anechoic chambers are essential [1] to the radiated EMI measurement. However, given its high cost and large space requirement, anechoic chambers and spectrum analyzers are usually not available in most power electronics labs. Also even with the anechoic chambers, the measured electric field intensity has limited reference values to the analysis and diagnosis of the common mode (CM) noise generated from the power converters.

For isolated power converters with long attached cables, the CM current along the attached cables can be a predominant contributor to the radiated EMI [2]–[9]. Measuring CM current with an oscilloscope is, therefore, a good way for the radiated EMI diagnosis. The measurement techniques for CM currents, impedances, and voltages are therefore very important.

The parasitic coupling effects between the power converter cables and the probe’s coaxial cable of the measurement equipment is insignificant in low-frequency range [3], [10], [11]. In high-frequency range, on the other hand, the couplings become significant because of the reduced parasitic impedance between the power converter cables and the probe coaxial cables [12]. The couplings can induce displacement currents flowing through the probe’s coaxial cables causing measurement errors. In order to improve the measurement accuracy and repeatability, the measurement technique needs to be improved.

Similarly, for the measurement of the voltage difference between the primary ground (PGND) and the secondary ground (SGND) of the power converter, in the low-frequency range, it usually has high magnitude, so measurement resolution is not an issue. But in high-frequency range, since the high-frequency voltage has a very small magnitude, the resolution of the oscilloscope is extremely critical. The measurement technique should be improved to measure the low magnitude voltages at high frequencies. Furthermore, the input impedance of the probe needs to be considered because it has loading effects on the measurement of the high-frequency voltage difference between PGND and SGND.

This article is an improved version based on the conference version [4]. New contributions of this article include the following:

1) improved analysis for the coupling issues in the CM current and impedance measurements;
2) improved analysis for the voltage transformations of a switching transformer through its parasitics;
3) improved analysis on the measurement of the voltage difference between PGND and SGND.

In Sections II and III, this article will identify and analyze the effects of the couplings on the CM current and CM impedance.
can be characterized with a lumped impedance $Z_{CMConv}$. The input and output attached cables can be characterized with an antenna impedance $Z_{Antenna}$, which is composed of resistance $R_L$ representing cables’ power loss, resistance $R_f$ representing cables’ radiated power, and reactance $X_A$ representing the near field energy of the cables [14].

According to the CM noise model in Fig. 1(b), the CM current $I_{CM}$ on the cables is given by

$$I_{CM} = \frac{V_{CM}}{Z_{CMTrans} + Z_{CMConv} + Z_{Antenna}}$$

$$= \frac{V_{GNDs}}{Z_{CMConv} + Z_{Antenna}}.$$  \hspace{1cm} (1)

**B. Coupling Issues in CM Current Measurement**

CM current on cables is measured using a current clamp probe, as shown in Fig. 2(a). $Z_g$ is the grid impedance and $V_g$ is the equivalent grid voltage source. There are two parasitic coupling issues. First, there are the parasitic couplings between input/output cables and the outer conductor (shielding) of the coaxial cable of the current clamp probe. The couplings can be represented with the lumped parasitic coupling impedances $Z_{CPi}$ and $Z_{CPo}$, as shown in Fig. 2(a). $Z_{CPi}$ and $Z_{CPo}$, can be extracted using $S$-parameters [15]–[18], [23]. As an example, to extract $Z_{CPo}$, a two-port $S$-parameter matrix is first measured between the output cable and the outer conductor of the coaxial cable of the current probe using the setup in Fig. 2(b).

Based on the network theory, the measured $S$-parameter matrix can be converted to impedance matrix and based on the impedance matrix, a $\pi$ impedance network can be derived in Fig. 2(c). In Fig. 2(c), $Z_1$ represents the coupling impedance $Z_{CPo}$ between two ports. $Z_2$ and $Z_3$ are the equivalent impedances to the ground so they are not part of $Z_{CPo}$. Based on the network theory, $Z_{CPo} = Z_1$ is as follows:

$$Z_{CPo} = 2Z_0S_{11}S_{22} - \left(1 - S_{12}\right)\left(1 - S_{21}\right)$$

$$\frac{S_{12}S_{21} - \left(1 - S_{11}\right)\left(1 - S_{22}\right)}{}$$ \hspace{1cm} (2)

where $S_{11}$ and $S_{22}$ are reflection coefficients and $S_{12}$ and $S_{21}$ are transmission coefficients.

Similarly, $Z_{CPi}$ can also be extracted. The extracted impedances are shown in Fig. 2(d). It is shown that the impedances are from 70 $\Omega$ to 1.6 k$\Omega$ in the concerned frequency range from 30 to 100 MHz. The CM current can, therefore, flow through parasitic coupling impedances to the coaxial cable of the current clamp probe and cause measurement inaccuracy as analyzed in later sections.

The second parasitic coupling is the CM current path from input cable, to $Z_g$ of the AC grid, to the ground and back to the output cable, as shown in Fig. 2(a). This CM current path can also degrade the measurement accuracy and repeatability for radiated EMI.

**C. Techniques to Reduce the Negative Effect of Couplings**

For the first parasitic coupling, Fig. 3 shows the equivalent circuit for analysis. $R_in = 50$ $\Omega$ is the input resistance of the measurement equipment. $Z_{CTout} = 5$ $\mu$H]$30$ pF is the output...
impedance of the current clamp probe R&S EZ-17 model 03. \(X_{r.g}\) and \(R_{r.g}\) represent the reactance and resistance of the coupling loop from the output cable, to ground, to grid impedance and to the input cable as discussed previously. The flyback converter’s CM noise model is in the box. The current clamp probe’s coaxial cable is modeled with two fully coupled inductances \(L_{Sh}\) and \(L_{CT}\) and two resistances \(R_{sh}\) and \(R_{CT}\) for the outer and inner conductors, respectively. The coupling currents \(I_{CPo}\) and \(I_{CPi}\), due to the voltage \(V_o\) on the output cable, the voltage \(V_i\) on the input cable, and the coupling impedances \(Z_{CPo}\) and \(Z_{CPi}\), which were measured in Fig. 2, flow to the outer conductor of probe coaxial cable. Due to the full magnetic coupling between the inner and outer conductors of the coaxial cable [19], the induced voltages on two inductances \(L_{Sh}\) and \(L_{CT}\) cancel each other in the input DM loop of the current probe. The measured error voltage \(\Delta V\) on \(R_{in}\) due to \(Z_{CPo}\) and \(Z_{CPi}\) is derived based on Fig. 3.

This coupling effect can be greatly reduced by adding ferrite beads to the coaxial cable. The ferrite beads should have high resistance in the high-frequency range, so they can greatly increase \(R_{CT}\) and \(R_{sh}\). The ferrite beads will not influence the measured current signal because it is a resistance to CM currents and the measured signal is a DM current in the probe. In the experiments, each ferrite bead Fair-Rite 0431167281 has a resistance 110–330 \(\Omega\) from 30 to 100 MHz. Ten ferrite beads with a total more than 1.1 k\(\Omega\) resistance are used. Based on Fig. 3 and (3), shown at the bottom of the the next page, \(\Delta V\) can be greatly reduced.

For the second parasitic coupling, Fig. 4 shows the equivalent circuit for analysis, where, \(Z_{CPg} = jX_{r.g} + R_{r.g}\). Due to the uncertainty of the grid impedance, \(I_{CM}\) varies at different grid conditions, so measurement is not repeatable. It is, therefore, necessary to reduce the coupling effect between the converter’s cables and the ac grid. To solve this, ferrite beads can be used on the cable between the converter’s input cable and the power grid.
In summary, to reduce the parasitic coupling between the power converter’s input and output cables and the shielding of the current clamp probe’s coaxial cable, ferrite beads can be used on the coaxial cable to reduce the CM current flowing to the coaxial cable; to reduce the parasitic coupling between the power cables of the converter via the power grid, the ferrite beads can be used on the cable between the input cable of the power converter and the power grid. Fig. 5(a) shows the implementation of the ferrite beads. Fig. 5(b) shows the comparison of the voltage spectra measured on the oscilloscope with and without the ferrite beads when the current probe is not clamped to the output cable. During the measurement, the converter worked as normal and the ferrite beads are used on the cable between the input cable of the power converter and the power grid. The positions of the current probe, the coaxial cable of the current probe and the converter’s cables are exactly the same as those during the normal CM current measurements. Theoretically, the measured voltage should be zero since there is no current flowing through the current clamp. However, as shown in Fig. 5(b), up to 60 dB \( \mu \)V voltage is measured on the oscilloscope. This verified that the coupling is not ignorable. After using ferrite beads, the spectrum is reduced by up to 40 dB, so the ferrite beads greatly help to reduce the undesired coupling to the coaxial cable. It is shown that \( Z_{CP_i} \) and \( Z_{CP_o} \) can cause significant error.

III. CM IMPEDANCE MEASUREMENT

From Fig. 1(b), for the model of the flyback converter, there are three CM impedances, \( Z_{CM\text{Trans}}, Z_{CM\text{Conv}}, \) and \( Z_{\text{Antenna}} \).

A. Extract Transformer Model

The windings of the transformer in Fig. 1(a) has a nonideal coupling coefficient, so the voltages of primary and secondary windings are not proportional to turns ratio at high frequencies. Furthermore, the transformer is a four-terminal device but its equivalent CM impedance \( Z_{CM\text{Trans}} \) is a two-terminal impedance, so a technique to convert a four-terminal device to a two-terminal impedance for CM current analysis should be investigated.

\[
\Delta V \approx \frac{(Z_{CP_i}V_o + Z_{CP_o}V_i)R_{in}}{Z_{CP_i}Z_{CP_o}(1 + \frac{R_{CT}}{R_{in}} + \frac{R_{in} + Z_{CT\text{Trans}}}{R_{in}}) + (Z_{CP_i} + Z_{CP_o})(R_{CT} + R_{in} + Z_{CT\text{out}})} \tag{3}
\]
In Fig. 1(a), based on substitution theory, both the primary and secondary MOSFETs can be replaced with voltage sources \( V_{pri} \) and \( V_{sec} \) which have the exact same waveform as those on the MOSFETs to be replaced in Fig. 6. The impedances of the dc bus capacitors on both input and output dc bus can be ignored for EMI analysis so they can be treated as short circuit. Based on the superposition theory, the CM noise generated from \( V_{pri} \) and \( V_{sec} \) can be analyzed individually by shorting the other voltage source. Fig. 7 shows the analysis of the contribution from \( V_{pri} \).

In Fig. 7, \( V_{pri} \) is the DM voltage added to the primary winding of the transformer. Due to the parasitic impedances between two windings, this DM voltage can generate \( V_{GNDs \_pri} \) across PGND and SGND [20], [21]. The voltage across PGND and SGND is the noise voltage added between the input and output cables. Because of this, there is CM current flowing through the input and output cables which behave like an antenna radiating EMI.

The parasitic inductance and capacitance between the two windings are mostly linear, it is possible to find the voltage transformation from \( V_{pri} \) to \( V_{GNDs \_pri} \) by first measuring \( S \)-parameter matrix in Fig. 8 and then convert it to voltage gains [17], [25]. In Fig. 8, ports 1 and 2 with input and output impedance \( Z_0 = 50 \Omega \) of a VNA are connected to primary winding and across the PGND and SGND, respectively. It should be pointed out that the setup in Fig. 2(b) is used to extract parasitic coupling impedances, whereas the setup in Fig. 8 is used to extract the circuit model of the transformer.

Based on the measured \( S \)-parameter matrix, a lumped \( \pi \) impedance model [17], [25] \( Z_1, Z_2, \) and \( Z_3 \) can be derived in Fig. 9(a). \( V_{GNDs \_pri} \) is the voltage on \( Z_3 \). Because \( Z_2 \) is in parallel with \( V_{pri} \), it can be removed. The transformer model can be further reduced to Fig. 9(b).

From Fig. 9(b), the CM transformation gain (CMTG\(_{pri}\)) from \( V_{pri} \) to \( V_{GNDs \_pri} \) can be derived as

\[
CMTG_{pri} = \frac{V_{GNDs \_pri}}{V_{pri}} = \frac{Z_3}{Z_1 + Z_3}. \tag{4}
\]

Similarly, the transformation CMTG\(_{sec}\) from \( V_{sec} \) to the \( V_{GNDs \_sec} \) can be analyzed and measured in Fig. 10(a) and (b). The total \( V_{GNDs} \) across the PGND and SGND is

\[
V_{GNDs} = V_{pri} \cdot CMTG_{pri} + V_{sec} \cdot CMTG_{sec}. \tag{5}
\]

Although both \( V_{pri} \) and \( V_{sec} \) contribute to the voltage \( V_{GNDs} \) across the PGND and SGND in (5), due to the magnitude difference between \( V_{pri} \cdot CMTG_{pri} \) and \( V_{sec} \cdot CMTG_{sec} \), one could be dominant. Fig. 11 shows the \( V_{pri} \cdot CMTG_{pri} \) and \( V_{sec} \cdot CMTG_{sec} \) based on the measured \( V_{pri}, V_{sec}, CMTG_{pri}, \) and \( CMTG_{sec} \). It is shown that for the flyback converter under investigation, \( V_{pri} \cdot CMTG_{pri} \) is dominant in the whole concerned frequency range from 30 to 100 MHz.

Because of this, based on Thevenin equivalence, Fig. 9(b) can be represented with (6) and (7) and Fig. 12.

\[
Z_{CMTrans} = Z_1 || Z_3 \tag{6}
\]

\[
V_{CM} = V_{pri} \cdot \frac{Z_3}{Z_1 + Z_3}. \tag{7}
\]

By the way, if both \( V_{pri} \cdot CMTG_{pri} \) and \( V_{sec} \cdot CMTG_{sec} \) significantly contribute to the voltage \( V_{GNDs} \) across the PGND...
B. Extract $Z_{\text{CMConv}} + Z_{\text{Antenna}}$

The total CM impedance of $Z_{\text{CMConv}} + Z_{\text{Antenna}}$ in Fig. 1(b) can be measured as shown in Fig. 13 after the transformer has been removed from the converter circuit. $Z_{\text{CMConv}} + Z_{\text{Antenna}}$ can be extracted using the two-port $S$-parameter as used in Fig. 2. The impedance can also be measured after shorting the two terminals of the primary as well as the two terminals of the secondary but the result is the same as that in Fig. 13 because of the small impedances of two dc bus capacitors.

Fig. 14 shows the extracted $Z_{\text{CMTrans}}$ and the measured $Z_{\text{CMConv}} + Z_{\text{Antenna}}$. The comparison indicates that $Z_{\text{CMTrans}}$ is smaller than $Z_{\text{CMConv}} + Z_{\text{Antenna}}$. It is, therefore, concluded that although the transformer is a critical component for $V_{\text{CM}}$ which finally drives the cables for the radiation, its CM impedance is a minor factor for the radiated EMI. This is especially true for compact transformers with big interwinding parasitic capacitances and small CM impedances.

IV. MEASUREMENT OF VOLTAGE ACROSS PGND AND SGND

The voltage $V_{\text{GNDs}}$ across PGND and SGND drives $Z_{\text{CMConv}} + Z_{\text{Antenna}}$ to generate the CM current and the radiated EMI. However, it also includes line-frequency 50 Hz/60 Hz voltage component. It was shown in Fig. 9(b) that the transformer can be modeled with two impedances across the primary and secondary sides. At low frequencies, these two impedances are mostly capacitance [20], [21] and the total of these two capacitances is equal to the total capacitance, which was measured as 24 pF, between the primary and the secondary windings of the transformer. Thevenin equivalent impedance $Z_{\text{CMTrans}}$ in Fig. 12 is, therefore, two capacitances in parallel with a total of 24 pF. There is also a parasitic capacitance between
the output dc bus, output cable and the earth ground. It was measured as 7 pF. Because the operating of the diode bridge, the dc bus on the primary side has a 50 Hz/60 Hz voltage referenced to the earth ground. Because $Z_{CM\text{Trans}}$, which is equal to 24 pF at low frequencies, is in series with the 7 pF parasitic capacitance between the output cable and the ground, there is a 50 Hz/60 Hz voltage drop between PGND and SGND. Depending on the values of ac voltage and parasitic capacitances, this voltage drop could be much bigger than the magnitudes of the high-frequency switching harmonics. For example, if the input voltage is 240 V/ac, with the capacitance above, $V_{GND}$ will have a 77 V 50 Hz/60 Hz peak voltage. The oscilloscope cannot easily differentiate high-frequency voltage with small magnitude, for example several mV, from the 50 Hz/60 Hz voltage with a big magnitude. For example, an oscilloscope with 8 digits data resolution has a resolution as poor as 0.625 V at 20 V/div scale. So, it cannot accurately measure mV level $V_{GND}$ and cannot catch the noise information in a full line cycle.

### A. Conventional Approach

Conventionally, an RC high-pass filter is used in Fig. 15 between $V_{GND}$, and the voltage probe to block 50 Hz/60 Hz voltage. The RC high-pass filter has a slope of 20 dB/dec. To efficiently reduce the 50 Hz/60 Hz voltage, the cutoff frequency $f_c$ of the filter can be chosen 5 dec higher than the line frequency. It is 5 MHz or 6 MHz.

If the high-pass filter is connected between the $V_{GND}$ and the voltage probe as in Fig. 15, the equivalent circuit is shown in Fig. 16. The input impedance $Z_{\text{Probe}}$ of the voltage probe is $C_{\text{Probe}} || R_{\text{Probe}}$. The input impedance $Z_{in}$ of the high-pass filter is given as

$$Z_{in} = R || Z_{\text{Probe}} + \frac{1}{j2\pi fC}. \quad (8)$$

### B. Proposed Approach

To solve this issue, the high-pass filter can be moved to between the voltage probe and the coaxial cable as in Fig. 18. Similarly, the condition (11) should be met to reduce oscilloscope’s loading effect on the voltage probe

$$|Z_{out}| \ll \left| \frac{1}{j2\pi f C_{OS} + \frac{1}{R_{OS}}} \right| \quad (11)$$

where $C_{OS} || R_{OS}$ is the input impedance of the oscilloscope. For Rigol MO4054, $C_{OS} = 15$ pF, and $R_{OS} = 1$ MΩ. Similarly, Considering $Z_{out} = R$ at the worst scenario, based on (15) at 100 MHz, $R \leq 106.16$ Ω. Also, the output impedance of the voltage probe is a 195 pF capacitance and it has an impedance

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**Fig. 15.** Conventional approach: a high-pass filter is implemented to block line-frequency voltage for $V_{GND}$ measurement.

**Fig. 16.** Equivalent circuit for $V_{GND}$ measurement with a high pass filter between $V_{GND}$ and the voltage probe.

**Fig. 17.** Comparison of $(Z_{CM\text{Conv}} + Z_{Antenna}) || Z_{CM\text{Trans}}$ and $Z_{in}$.

**Fig. 18.** Improved implementation of the high-pass filter.
smaller than 27 Ω above 30 MHz, so $R$ should be much larger than 27 Ω. Based on these conditions, the RC filter is designed as $R = 100$ Ω, $C = 300$ pF. For the same reason as the current probe, to reduce the parasitic coupling to the coaxial cable of the voltage probe, ferrite beads are used on the coaxial cable before and after the high-pass filter. Furthermore, because the voltage probe has an input impedance of $Z_{\text{Probe}} = 13$ pF || 1 MΩ, it has a loading effect on the $V_{\text{GNDs}}$ measurement as it is equivalently in parallel with $Z_{\text{CMConv}} + Z_{\text{Antenna}}$. So the directly measured $V_{\text{GNDs}}$ is not the actual $V_{\text{GNDs}}$ when the voltage probe is not connected between PGND and SGND. To solve this issue, high resistive ferrite beads are used on both input and output cables of the converter to provide high impedance to block the antenna impedance, so the measured $V_{\text{GNDs}}$ will be the output of the voltage divider composed of $Z_{\text{CMTrans}}$ and $Z_{\text{Probe}}$. The $V_{\text{CM}}$ and actual $V_{\text{GNDs}}$ can, therefore, be derived based on the measured $V'_{\text{GNDs}}$ as

$$V_{\text{CM}} = V'_{\text{GNDs}} (1 + Z_{\text{CMTrans}}/ Z_{\text{Probe}})$$

$$V_{\text{GNDs}} = V_{\text{CM}} (Z_{\text{CMConv}} + Z_{\text{Antenna}}) / (Z_{\text{CMConv}} + Z_{\text{Antenna}} + Z_{\text{CMTrans}}).$$

V. EXPERIMENTAL VERIFICATION

Fig. 19 shows the measurement setup based on the proposed measurement techniques on a flyback converter on a table. The input voltage of the converter is 240 V/60 Hz, the output is 5 V/4.5 A. The switching frequency is 65 kHz. The output cable is a 0.9 m USB-A to USB type-C cable with a load adapter and a 1 Ω load resistor. The height of the table is 0.8 m. The input cable is perpendicular to the ground as shown in the figure. High-frequency resistive ferrite beads (Fair-Rite 0431167281) are used on the coaxial cables of the current clamp probe and the voltage probe. The ferrite beads are also used on the cable on the ground between the input cable of the flyback converter and the ac grid. The oscilloscope is Rigol MSO4054. The current probe is EZ-17 Model03 from Rohde & Schwarz and the voltage probe is Rigol RA3500A. In $S$-parameter measurement, the VNA is Planar 808/1 from Copper Mountain.

Fig. 20 shows the spectrum comparison of the $V_{\text{CM}}$ derived from the directly measured $V'_{\text{GNDs}}$ based on the measured time domain waveforms with and without using the proposed high-pass filter.

The background noise with the proposed high-pass filter when the converter is turned off is also measured in Fig. 20. It is shown that without using the proposed high-pass filter technique, because the resolution of the oscilloscope is not enough for high-frequency low magnitude voltage measurement, the whole spectrum shows high noise floor in Fig. 20. On the other hand, the background noise with the proposed high-pass filter is much smaller than the derived $V_{\text{CM}}$. Because of this, the proposed high-pass filter improved the signal-to-noise ratio.

With the $V_{\text{GNDs}}$ and $Z_{\text{CMConv}} + Z_{\text{Antenna}}$, the CM current on the input and output cables can be predicted based on Fig. 4, as shown in Fig. 21. On the other hand, based on the directly measured CM current time-domain waveforms, the spectrum of the CM current is also derived in the figure. The dashed black curve is the CM current spectrum derived from the directly measured time-domain waveform of the CM current without using any ferrite beads. In Fig. 21, the difference between the predicted and the measured CM current spectra is less than 6 dB. Therefore, the proposed techniques for CM current, impedance, and voltage measurement are verified. It should be pointed out that the measured CM currents on input and output cable are almost the same. The CM current is also almost constant along the cables. The spectrum of the measured CM current without using any ferrite beads much deviates from the accurate value, so it is incorrect.
VI. CONCLUSION

In this article, several measurement techniques are proposed for CM current, voltage, and impedance measurement for radiated EMI analysis. The coupling between the coaxial cable of a current probe and the input and output cables of a flyback converter is analyzed. The model is developed and a technical solution is proposed. The CM model of a switching transformer is developed and its voltage transformations from the DM voltages to the voltage across PGND and SGND and its impedances are extracted based on measurements. A high-pass filter implementation for measuring the voltage across PGND and SGND is proposed. The proposed techniques are finally validated in experiments.

REFERENCES


