

# Compact Laser Driver for Ultrasonic Arbitrary Position and Width Pulse Sequences Generation

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**Abstract**—Laser ultrasound offers several benefits in comparison to immersion or air coupled ultrasound: it directly generates stress on the sample surface; it has a wide bandwidth and a small acoustic source can be produced. Conventionally, high power pulsed lasers are required therefore equipment is bulky and expensive. Here we propose to use mid-power laser diodes, which are small and low cost. To solve the low signal amplitude problem with lower power lasers, the use of spread spectrum signal, arbitrary position and width binary pulse sets is proposed. To date, a laser driver for this task has not been available. This paper describes the development of such a compact driver, where the essential electronics occupy an area of 10x20 mm. The whole system with a fixture for kinematic mount is comparable in size to a conventional piezoelectric ultrasonic transducer. The driver can supply pulse sets of up to 40 A. Individual pulse duration can vary between 20 ns to 1000 ns (0.5-25 MHz ultrasound range) and the total pulse train duration is limited by the laser type used. Experiments show that up to 10  $\mu$ s long pulse sets can be used at 10 A current, without laser degradation. Current waveforms, beam profile and optical response signals were measured for three rectified topologies. It was concluded, that the GaN based constant current switch topology has the best performance, but a power MOSFET in a source current feedback topology can also be used to generate pulses down to 20 ns. Photoacoustic response signals from chirp and phase shift keying modulation have been demonstrated.

**Index Terms**—Arbitrary position and width pulse sequences, acoustic point source, diode lasers, laser ultrasound, laser driver, thermoelastic regime, ultrasonic transducers.

## I. INTRODUCTION

ULTRASONIC measurement and imaging are used widely across a range of application areas. One of the outstanding ultrasonic equipment features can be the small size, low cost of transducers and the absence of ionizing radiation [1]. However,

the bandwidth is limited if a conventional, piezoelectric transducers are used [2]–[4]. Unfortunately, the efficient energy transmission into the test media requires a piezoelectric transducer to be in contact with the test medium. Noncontact transduction is possible using air-coupled ultrasound which can be challenging with high impedance mismatches, or electromagnetic acoustic transducers (EMATs) which whilst being non-contact still must be within a few millimetres of the sample surface [5]–[7].

Laser ultrasound, which was first proposed 60 years ago [8] offers several benefits in comparison to immersion or air coupled ultrasound [9]–[13] for some applications. First of all, it is directly coupled to the surface of the test sample so: i) acoustic contact is stable [14] and ii) components that are sensitive to water can be inspected [15]; iii) impedance mismatch does not affect the signal level [11]; iv) reverberation in coupling media as observed with some contacting transducers is avoided; iv) accurate transducer focal point and test sample alignment is not necessary [16]. Other advantages are that acoustic source is not resonant, so that wide bandwidth is possible [10]; the distance from the light source to the test sample can be large [13]; arbitrary shapes of the acoustic source can be provided using special optics [17]–[22]; and the probing point can be confirmed visually (if a visible light laser is used or via an IR camera).

Three regimes of ultrasound production via laser excitation can be identified [9]: i) plasma or ablation regime uses  $\geq 10^7$  W/cm<sup>2</sup> (for metal) [13], [23]; ii) thermoelastic  $< 10^7$  W/cm<sup>2</sup> and iii) intermediate [24].

Thermal expansion of the volume, heated by laser pulse is the major cause of the ultrasound production in the thermoelastic regime, where incident power densities are below the damage threshold of the test material [25]–[27]. The amplitude of the acoustic waves produced (longitudinal, shear and Rayleigh) is proportional to the energy absorbed from the laser beam [25]–[28].

More mechanisms produce ultrasonic waves in plasma regime: light pressure, thermal expansion, electrostriction, and material ablation recoil [9],[10],[29]. Very high amplitude signals, into nonlinear ranges of ultrasound, can be obtained

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[10],[30]. It is sometimes essential that a longitudinal wave propagating normal to the surface is produced, but the resultant wave can have complex and wide beam directivity. Laser ablation can also change the material surface, so subsequent signals are different if obtained in the same location [9] and material damage occurs, particularly where coating or surface layers with different properties to the bulk are removed by the first laser beam shot.

Another problem, related to the thermoelastic regime laser ultrasound is the production of dual acoustic pulse when the free surface thermoelastic source is used. Transient heating of the material surface produces two pulses, which travel in opposite directions: one towards the material bulk, another-outwards. This second pulse is reflected back into material. The distance between the pulses is small, usually just the laser penetration depth (5-10 nm for metals [10]). In such case, the second pulse almost immediately (fractions of nanosecond) cancels the first pulse. Therefore, the generation of longitudinal waves normal to the material surface is complicated in the thermoelastic regime. There are applications that make use of the angled or surface waves [14],[31]-[35], and others that require the main beam to be directed normal to the surface. In these cases a constrained (buried) laser ultrasound source can be considered (sometimes addressed as fourth regime in laser ultrasound). In this case the surface is either coated with an optically transparent layer on top of the sample, resulting in constrained acoustic source [36]-[45]. Then second pulse is either delayed or cancelled, giving rise to the longitudinal wave propagating normal to the surface being enhanced by several orders when compared to the unconstrained surface. When a transparent coating is applied to the surface of the probed material, a second ultrasonic pulse travels in the coating until it is reflected back from the air interface. If the coating thickness is comparable to the ultrasound wavelength and the attenuation in coating is low, then resonant effects can be observed. The presence of the coating also reduces the amplitude of the surface wave, which is usually induced in unconstrained surface [41]. A constrained surface is considered in this paper.

When excited by a single pulse, the resulting signal lacks high frequencies and there are spectral dips because of the *sinc* shape of the spectrum. Excitation using the spike somehow leverages the problem, but the amount of energy per bandwidth of interest is very low. The wider is the required bandwidth, the higher is the desired frequency for excitation, the shorter should be the pulse duration. A large excitation power then has to be used to counter the lack of energy. This in turn calls for the bulky and expensive laser source. Then the essential ultrasound advantages like safety, small size and low cost are lost.

Spread spectrum (SS) excitation can solve the challenge of using a generation source with low pulse energy because bandwidth is not related to the pulse duration for SS signals [3]. Whilst widely used in conventional ultrasonic techniques, the SS application for use with laser ultrasound is complicated.

A CW laser (100 mW to 2 W power) can be modulated by a SS signal, which after reception can be compressed in software [46]-[52]. However, the power transmitted is very low, and the thermoelastic wave produced is small, so significant averaging

of very long spread spectrum signals has to be used; the modulation is shallow and all the equipment is analog, so that efficiency suffers.

There is an intermediate solution, which has all the conditions required for successful exploitation in laser generated ultrasound: mid-power ultrasound, using binary sequence excitation. Laser diodes of 10 W or even 600 W pulse power, which are now off the shelf components. Pulse durations in the range of 10 ns to 1000 ns are possible, which fits well onto the range required for 0.5-50 MHz ultrasound generation. The complexity of the optics required is reduced, thanks to the wide availability of relatively low cost aspheric lenses. The only missing element required to create a small laser ultrasound source is the laser driver. If all the components can fit into the size of the conventional ultrasonic transducer (approximately 50 mm long, 20 mm diameter) and at a comparable price, then the development of mid-power laser ultrasound will lead to many applications. There are many applications for such medium power lasers [53]-[63], mainly in photoacoustic imaging, where only a single pulse excitation is usually used.

There have been previous attempts to use spread spectrum excitation using mid-power (100 W) pulsed laser [64], but researchers had to use emulation, due to the unavailability of a fast, high current laser driver for generating pulse sequences. The acoustic response to a single pulse was measured and then pulse sequences were produced artificially, emulating spread spectrum sequence excitation.

In this work, we present the development of the complete laser ultrasound source. Targeted applications are the synthetic aperture focusing technique (SAFT) [10],[65]-[68], photoacoustics [53]-[63] and the transducer spatial sensitivity evaluation and calibration [69]-[72]. The whole laser system is compact and lightweight. It is capable of generating a small (down to 100  $\mu\text{m}$ ) acoustic source of arbitrary position and width pulse (APWP) sequences [69] (i.e. not a single pulse but multiple pulse trains).

The majority of the research effort was concentrated on developing the driver electronics. The beam shape control was limited to a simple optics, aiming for simplicity and small size. The experimental investigation of the ultrasound generation on an acoustic delay line is presented, demonstrating that the spectral content of the ultrasonic signal can be controlled using the APWP sequences, described in [69] and [73].

## II. REQUIREMENTS STUDY

There are certain limitations that have to be considered when efficient ultrasound excitation in the thermoelastic mode is used: i) the limited available power and ii) the requirement for a wide range of pulse durations and their repetition frequency.

### A. Attainable Power Density

Power density used in high power ultrasound usually is close to the ablation threshold,  $10^5$ - $10^7$  W/cm<sup>2</sup> [9],[10], [31]-[45]. The peak power has to be from tens of kW (42 kW in [53]), to hundreds of kW ([74], 833 kW in [43], 555 kW in [41], 300 kW in [19], 384 kW in [21]) or even MW (1.25 MW in [75], 1.7 MW in [36]). Usually researchers aim to use large (mm

order) laser spot because this increases longitudinal waves energy directed normal to the surface, but power densities remain in MW/cm<sup>2</sup> range.

Laser diodes used in this study are mid-power (refer Table I: SPL UL90AT08, SPL PL90\_3 are from OSRAM Opto Semiconductors GmbH, Regensburg, Germany and 905D1S3J09UA from Laser Components GmbH, Germany).

TABLE I  
LASER DIODES USED IN THE INVESTIGATION

Model	P <sub>max</sub> , W	Aperture, μm	I <sub>max</sub> , A	V <sub>F</sub> , V	Θ <sub>FWHM</sub> , deg
SPL PL90_3	75	200x10	40	9	9/25
SPL UL90AT08	120	220x10	40	11	10/25
905D1S3J09UA	80	235x10	35	12.8	12/20

It is difficult to reach the aforementioned power density with a laser diode whose power is limited to 75-120 W. Assuming a 5 mm spot size and 120 W laser, would result in a power density of 600 W/cm<sup>2</sup>. However, we see the merit of the small spot size. If a 120 W beam is focused into a 200 μm diameter spot, the power density would be 380 kW/cm<sup>2</sup>. Alternatively, a 75 W laser at a 100 μm spot diameter can produce 10<sup>6</sup> W/cm<sup>2</sup>, which is close to the power densities that provide sufficient acoustic output (assuming 100% absorption). Many application can benefit from a small spot size: i) a wide acoustic beam is required in SAFT imaging [65]-[68]; ii) if the material investigated is thin, then the acoustic beam can be considered already focused so high lateral resolution is attained [70]; iii) photoacoustic imaging requires small laser beam size [53]-[63]. However, laser beam spot size cannot be reduced indefinitely, and is usually limited by the emitting aperture size (10 μm x 200 μm is common) and the optical system magnification/reduction. The small laser beam spot size is supposed to result in a small acoustic generation spot size, but it could be larger [39].

### B. APWP Sequence Parameters

Replacing the single pulse excitation with APWP sequences offers several benefits. Long signals can be used, so signal energy is boosted not by increasing the instantaneous power (pulse amplitude) but by signal duration. Application of a matched filter (cross-correlation function) allows one to compress the signal, obtaining a sharp pulse. The signal spectrum can be designed to match the required bandwidth, so that all of the energy is used to produce useful frequencies. Furthermore, the spectral content of the ultrasonic signal can be programmed [73]. APWP sequences can be considered as a binary analog of an arbitrary waveform. In this paper, only laser On or Off states are available. According to signal theory, the optimal pulse duration is half of the period of the desired frequency component. When the conventional ultrasound range of 0.5-20 MHz is considered, the pulse durations should be from 25 ns to 1000 ns. Several well-known signals can be generated using APWP sequences: single pulse, CW toneburst, phase shift keying (PSK) sequences, chirp or completely arbitrary pulse sequences. CW toneburst can have a fixed duty cycle and pulse repetition period equal to the desired frequency

$f_0$ , or a variable duty cycle to adjust the amplitude or impose an envelope modulation. For PSK sequences, the chip (the elemental atom of a PSK sequence) is constructed following the toneburst principles: the pulse duration defines the bandwidth, and the repetition frequency defines the center frequency. Chirp signals can be made using the linear or nonlinear frequency modulation (LFM or NLFM), it can have a tapered amplitude by using duty cycle variation [73]. Arbitrary pulse sequences can adapt to the required spectral [73] or correlation properties [69] by optimization. The total sequence length is limited by the laser diode's safe operating conditions: at 40 A it should be limited to around 500 ns, but for 10 A APWP sequence length can reach tens of μs.

### III. LASER DRIVER DEVELOPMENT

Laser diodes are usually low voltage devices, where the forward voltage is 5-12 V, even for stacked lasers. Therefore, a high current of up to 40 A is required to deliver the power required. Laser drivers that are commercially available are large and are not intended for relatively short duration and high intersymbol rate APWP sequences generation.

The laser driver PCO-7114-22-2 is available from DEI (Directed Energy, Inc., San Rafael, California USA), and is relatively small (62 x 38 mm), producing up to 22 A pulses, but can only produce a single pulse of a fixed <2ns duration. The PCO-7110 driver can produce a 4-65 ns, 4 A to 120 A pulse, but is not capable of driving sequences (pulse repetition frequency (PRF) is 50 kHz). The PCO-7120 (51 x 103 x 19 mm) can produce pulse widths of 12-1000 ns at 5 A to 50 A per pulse, but the PRF is limited to 1 MHz. The laser driver BFS-VRM 03 offered by Laser Components (Laser Components GmbH, Olching, Germany) can produce a wide range of pulse width of 1 ns to CW output, with a size of 65 x 85 mm, yet the maximum output current is just 3 A. The same company offers the pulsed driver LSP-40, which can deliver 40 A, 30-1000 ns single pulse width (PRF is 10 kHz). Hamamatsu (Shidzuoka, Japan) offer a laser driver board C14518, that can deliver up to a 40 A pulse, but the single pulse duration is fixed to 4 ns, and the PRF is limited to 150 kHz). Quantel have reported a laser driver capable of 1.7 mJ, but again, the pulse output is single and a fixed 40 ns duration [53]. To summarize, commercially available drivers offer just a single pulse and are mostly aimed for short, <10 ns, pulse generation.

Short laser pulses are of interest in lidar, and there has been significant effort in laser drivers development for this purpose [76], [77], [78]. Yet, the solutions proposed are either too low power and pulse durations are not within the range of interest discussed here, or the pulse repetition frequency is too low. Section III.C provides a more detailed discussion on topologies.

To conclude, conventional pulsed laser drivers are aimed at generating single short pulses, but the application presented in this paper requires wide range, of pulse width of up to 1000 ns duration and repetition frequency close to 30 MHz. The driving current should reach its "on" state as fast as possible, and should be maintained over the whole duration of the pulse.

### A. Main Switch FET

The selection of the FETs used in the main switch was based on maximum current (40 A from table I) and maximum speed. A smaller reverse capacitance  $C_R$  reduces the parasitic feedback (oscillations). Total gate charge  $Q_{Gtot}$  (together with gate driver current  $I_{out}$  capability) and gate capacitance  $C_G$  define the turn-on speed. The FET output (Drain-Source) capacitance  $C_{outMax}$  at 0 V Drain-Source voltage, influences the switching losses, oscillation and edge duration. The drain-Source breakdown voltage  $V_{DSmax}$ , defines the attainable compliance voltage (40-50 V were required; see section C). Two FET types were selected (refer Table II): the most popular enhancement mode GaN (EPC2019 in 2.7x0.9 mm wafer level chip scale package (WLCSP) package, from Efficient Power Conversion Corporation, El Segundo, CA, USA) and a power MOSFET (NVTFS6H888N in WDFN8 3x3 mm package from ON Semiconductor, Phoenix, AZ, USA; BSZ340N08NS3, OptiMOS3 and BSZ146N10LS5, OptiMOS5 in 3x3 mm TSDSON-8 package from Infineon Technologies AG, Neubiberg, Germany).

TABLE II  
FETs USED IN INVESTIGATION

Model	$V_{DSmax}$ , V	$I_{Dmax}$ , A	$Q_{Gtot}$ , nC	$C_R$ , pF	$C_G$ , pF	$C_{outMax}$ , pF
EPC2019	200	42	1.8	0.7	200	380
NVTFS6H888N	80	47	4.7	3	220	320
BSZ340N08NS3	80	92	6.8	7	470	600
BSZ146N10LS5	100	176	15	9	1000	1000
BSZ900N15NS3	150	52	5	3	380	600

While GaN is the obvious choice, it has some disadvantages: i) the gate voltage range is very narrow, so that gate boosting might be dangerous (MOSFETs are more immune to gate transients, within the +/- 20 V range, while GaN have only -4 V to +6 V); ii) the WLCSP package is difficult to solder manually and to inspect for soldering quality. Besides, it was interesting to investigate what parameters power MOSFET can achieve.

### B. FET Driver

Four FET drivers were selected (refer Table III).

TABLE III  
FET DRIVERS USED IN INVESTIGATION

Model	$V_{out}$ , $V_{min}$	$I_{out}$ , A	$R_{out}$ , $\Omega$	$t_{dly}$ , ns	$t_r$ , ns
LMG1020	4.3-5.4	7/5	n.a.	2.5	0.37/0.35
UCC27511	4.2-18	4/8	5/0.45	10/12	7.5/5
MAX5048	4-14	7/3	0.88/0.32	8	5/4
ISL55110	0-14	3.5	3	10	1.2/1.4

The LMG1020 (0.8x1.2mm DSBGA package, from Texas Instruments Incorporated, Dallas, TX, USA) was chosen because of the low propagation delay time and impressive rise/fall times (it should be noted that these are specified at very light, 100 pF load). Yet, this driver has a serious limitation: the available output voltage range is very narrow. As will be seen in the following explanation, a regulated driver output voltage was required. The UCC27511 (SOT23 package, from Texas Instruments Incorporated, Dallas, TX, USA) and MAX5048

(SOT23 case, from Maxim Integrated, San Jose, CA, USA) drivers have similar performance, with one essential difference: MAX5048 can source up to 7 A and has much lower (0.88  $\Omega$ ) source impedance, while the UCC27511 is almost the opposite, having a high (8 A) sink current. The ISL55110 (TSSOP package, from Maxim Integrated, San Jose, CA, USA) is a special case, and is particularly suited to the application discussed in this paper: it has separate power supplies for the internal logic and the output stage. Therefore, it can provide very low output voltages without significant performance degradation (driver output FETs' driving voltage can stay fixed). It also has two channels in one package, which have been paralleled to get 1.5  $\Omega$  and 6 A output. Simulation using LT Spice did not reveal significant differences in the drivers' performance, and so it was decided to concentrate on two drivers: the ISL55110, with two channels paralleled and the LMG1020 only for GaN in switch mode.

### C. Parasitics

The inductance of the laser diode current loop, including the printed circuit board (PCB) [79], limits the rise time of the current injected. According to [80] the laser contributes most of the inductance, and the majority of the inductance comes from the laser package, including the wire bonds. The inductance value stated in [80] is 5 nH. Very short laser diode pins (2 mm long), were soldered directly on PCB to keep the laser inductance as low as possible. Our measurements indicate that in such a case, the inductance of the laser diode is 2.5 nH at 10 mA bias (measurements at higher currents were not done in order to avoid laser damage). Delivery of the high current pulses to a laser diode is also a challenge [80]: a microstrip type, low inductance transmission line has to be used, and this still introduces parasitics. Therefore, it was decided that the laser driver has to be integrated with the laser diode on the same PCB. The PCB trace also contributes significant inductance: a 1 mm wide, 10 mm long trace on 1.5 mm FR4 substrate would give 19 nH, so it was decided that wide traces and thin PCB would be used. The 1.5 mm PCB with 10 mm wide and 6 mm long trace would contribute 1.14 nH inductance, but an 0.8 mm PCB would have 0.47 nH inductance (0.9 nH if both forward and returning current path is considered). The MLCC capacitor, according to [81] has 870 pH, 1050 pH and 1200 pH for the 0603, 0805 and 1206 respectively, while the inverted 0612 has 610 pH and the 0508 has 600 pH. The inverted MLCC case was used in the main current loop. Resistors have even higher parasitic inductance if in a standard configuration (resistive layer on top of ceramic substrate): wrap-around effect causes the 0603 to have 1600 pH and the 0402 to have 120 pH inductance [82]. Fortunately, flipped versions (resistive layer at the bottom) have 26.7 pH for the 0603 and 2 pH for the 0402. Several components can be used in parallel, further minimizing the parasitic inductance, yet it has been reported that five parallel 0402 shunt resistors flipped, still have 40 pH inductance [80]. Inverted and flipped case resistors were used in the main current loop.

FET manufacturers usually do not specify inductance, but data was available for the WDFN8 case (one used in this

research): the drain inductance is 0.2 nH and the source inductance is 0.65 nH [75].

The expected loop inductance, even considering the modifications mentioned is 4.6 nH. The required compliance voltage can be calculated from this inductance. A simplified equation for the rising edge is:

$$t_R = \frac{dI \cdot L}{V}, \quad (1)$$

where  $dI$  is current step size,  $V$  is the excitation voltage,  $L$  is the loop inductance and  $t$  is time. If 25 ns is the shortest pulse to be produced, then  $t_R$  should be at least 5 ns. In such a case, a 40 A step requires 36 V. If the forward voltage of the laser diode is 9-12 V, then the compliance voltage should be within the 40-50 V range. It should be noted that this equation does not take into account that the MOSFET switch is a current sink at high  $V_{DS}$  voltages, and is a resistor at low voltages, so that actual rise times will be longer. Yet, there is an essential problem that can be seen from (1): a high compliance voltage is required in order to achieve fast rise time [83].

#### D. Driver Topologies Analysis

A circuit with resistor current limiter can be considered as the simplest one (Fig.1 left).

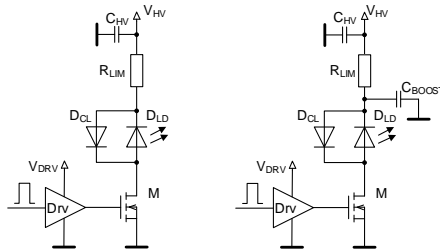


Fig. 1. Resistor current limiter topology (left) and boosted version (right).

Here, the laser diode current is defined by the voltage  $V_{HV}$ , laser diode forward voltage and limiting resistor  $R_{LIM}$ , plus the FET resistance  $R_{DS}$ . Moderate drive current stability can be achieved with a stable  $V_{HV}$  voltage and low FET temperature variation. The circuit is simple, and current monitoring can be done by measuring the voltage drop on the resistor  $R_{LIM}$ . A clamping diode  $D_{CL}$  is used for the FET  $M$  and laser diode  $D_{LD}$  protection, against the voltage transients caused by the cut-off of the high loop current.

This circuit has been simulated using LT Spice, using 0.5  $\Omega$   $R_{LIM}$  40 V  $V_{HV}$ , and a 1.5  $\Omega$  gate driver (equivalent to two ISL55110 channels in parallel). The results are presented in Fig.2.

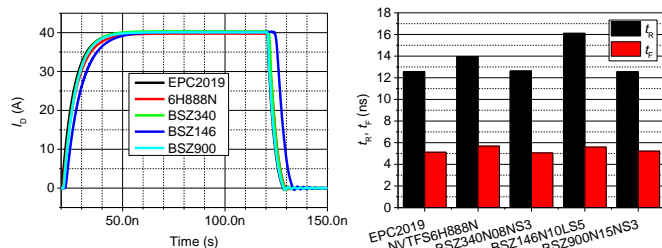


Fig. 2. Resistor current limiter topology driving current waveforms (left) and edge durations (right).

The simulation used only one laser diode model, which had highest forward voltage. It was derived from the RLD90QZW3 (Rohm Semiconductor) model, which has similar parameters to the 905D1S3J09UA model. Simulation results indicate that 40 V  $V_{HV}$  is insufficient: only 12.5 ns to 16 ns rise times were obtained. The reason is very simple: the compliance voltage is less than  $V_{HV}$  by voltage drop on  $R_{LIM}$  (20V) plus FET voltage (6 V for NVTFS6H888N and BSZ900N15NS3, 1.6 V for EPC2019, 1.2 V for BSZ340N08NS3 or 0.5 V for BSZ146N10LS5 at 40 A).

The circuit can be easily boosted to give more current during the rise time (Fig.1, right), and the boosted resistor current topology has been simulated using  $C_{boost}$  15 nF and  $V_{HV}$  40 V, results are in Fig.3.

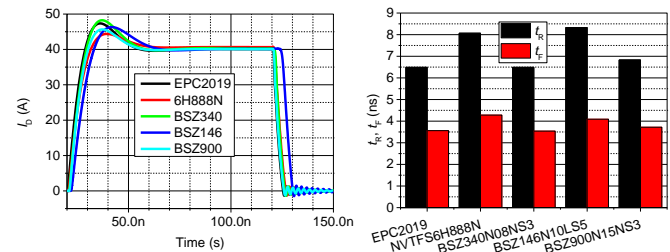


Fig. 3. Boosted resistor current limiter waveforms (left) and speed (right).

From this simulation, it can be concluded that the rising edge duration has been reduced significantly, down to 6 ns and all FETs have similar performance. Higher  $V_{HV}$  (and larger  $R_{LIM}$ ) would allow for further switching time reduction. The fact that the NVTFS6H888N performance is slightly worse than that for the BSZ900N15NS3 is unexpected, as these MOSFETs have similar parameters. Notable, that the falling edge times were also reduced (from 5 ns to 3.5 ns). The boosted resistor current limiter was selected as the candidate with GaN FET EPC2019.

The driver described in [84] is based on a standard capacitor discharge topology. It has a small footprint (50x50 mm), and can only reach 10 A with 1-1.5 ns single pulse widths at 100 kHz PRF, and a compliance voltage of up to 150 V is required. Circuit (Fig.4) can be modified as in [78], [80] so that its performance becomes similar to the boosted resistor current limiter topology (Fig.1 right), and only  $R_{lim}$  and  $C_1$  have different function and values here.

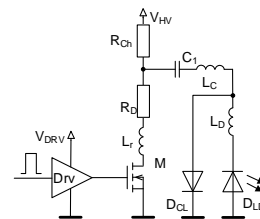


Fig. 4. Capacitor discharge/resonant laser diode drive topology.

Here, capacitor  $C_1$  is charged via the  $R_{ch} - D_{CL}$  network. Turning on, switching the  $M$  will cause the  $C_1$  to discharge into the laser diode  $D_{LD}$ . The circuit is usually made resonant so the resulting pulse width is [76]:

$$\tau_{0.5} \approx 2 \cdot \sqrt{L_{\Sigma} \cdot C_1}, \quad (2)$$

where  $L_{\Sigma}$  is the total loop inductance (usually formed just by

series connection of the PCB parasitics,  $L_\tau$  is the capacitor stray inductance and  $L_C$  is the inductance of the laser diode ( $L_D$ ).

The peak current of the laser diode is defined by  $V_{HV}$ ,  $C_1$  and  $L_\Sigma$  if circuit is optimally damped [76]:

$$I_{pk} \approx \frac{V_{HV}}{R_D \cdot \sqrt{\frac{L_\Sigma}{C_1}}} \quad (3)$$

A major disadvantage of such a circuit is that pulse duration and current depend on parasitics and the  $C_1$  parameters, so it is not suitable for producing a wide range of pulse durations. Furthermore, a large  $C_1$  is required for long pulses, making charging time very long, which is why this topology usually has only Hz to kHz range PRF. Another aim of the research described in this paper is to develop a circuit suitable for PRF of up to 30 MHz, within the sequence. The PRF can be increased using very low  $R_{Ch}$  values, but it will affect the resonant tank and severely increase the load on switch  $M$ .

The same solution is used in [85], where a GaN FET can deliver up to 30 A, 0.5-25 ns wide pulses at a PRF of up to 5 MHz, or 60 A, 50-500 ns wide single pulse as per [86]. Mode-locked lasers [87] can reach 100 MHz PRF, even at 20 A, but pulses are short (<2 ns) and the PRF is fixed: this solution is not suitable for APWP sequences. The laser driver presented in [85] uses GaN transistors, and can deliver 400 A, but only for a 50 ns wide single pulse.

An interesting solution was described in [89], using a delay line with a variable load (Fig.5). It can achieve current only up to 1.2 A, at a 20 MHz PRF and 200 ps up to 2 ns pulse widths.

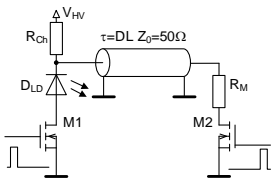


Fig. 5. Charge-line dual FET laser diode drive topology.

Here the delay line, charged via  $R_{Ch}$  together with switch  $M2$  is used to control the pulse duration. The current is mainly defined by the ratio of  $V_{HV}$  and  $Z_0$ , so that only low currents can be developed. Furthermore, the pulse width maximum duration is defined by the delay line delay value  $DL$ , so that a very long delay line would be needed, in order to achieve pulse widths of 1000 ns.

The inductive boost topology (Fig.6) uses an inductance  $L_{Ch}$  (could also be PCB trace), precharged by a constant current source  $I_{cc}$ , to immediately set the compliance voltage that is required to drive laser diode parasitic inductance. The laser diode is always tampered by  $M$ , and when a current pulse is required, the current accumulated in  $L_{Ch}$  is released into  $D_{LD}$ , to generate a fast rising edge.

A similar solution presented in [90] is very attractive and energy-efficient: a constant current is produced by a DC/DC converter, that can achieve 2.5 MHz PRF at 5 A, but requires three switches and complex driving signals, which should be implemented in dedicated IC or FPGA.

The avalanche bipolar junction transistors can produce high

power pulses, but pulse widths are short and the topology requires high voltage (300 V in [91]).

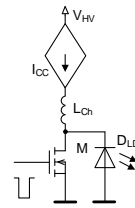


Fig. 6. Inductive boost laser diode drive topology.

An interesting capacitive boosting solution is described in [92], where a switched capacitance is used to provide extra compliance voltage during the current rise time. The solution described can only achieve pulses up to 1 A, and pulse duration is limited by the capacitance used. It requires six switches and complex driving logic. The circuit in Fig.1, right can be considered to be a simplified version of this idea. A capacitive charge transfer and voltage doubler combination is presented in [93], which can achieve impressive 330 MHz PRF, but current is limited to 180 mA, with a pulse width of 252 ps.

The laser driver presented in [94] uses an innovative resonant laser drive to deliver fixed a 20 A current at a PRF of up to 40 MHz, but the pulse width duration is short and fixed to 1 ns. Resonant driving is also used in [95] to achieve a current of 35 A and pulse widths below 4 ns, in comparison with BSZ146N10LS5, which is done by simulation. The topology inherently limits pulse duration and PRF. The laser driver presented in [96] is capable of a 100 MHz PRF, but the pulse width is fixed at 5 ns and has a peak current of only 2 A. The same resonant drive can achieve a current of 150 A for a fixed 8 ns wide single pulse [97]. The design described in [98] is an integrated solution, using the current sink topology, and can deliver up to 80 A, but only up to 25 ns single pulse. A similar approach [99] achieved a current of 10 A for a single 10-50 ns pulse. The low side switch solution presented in [100] obtains pulse widths down to 2.5 ns at currents of up to 9 A, with a 200 MHz PRF and pulses with 1 ns rise and fall times. A new concept can be derived from the topologies discussed: the FET can act both as a switch and as a current sink (Fig.7).

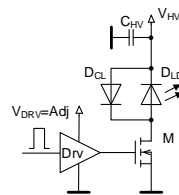


Fig. 7. FET acting as both switch and current sink laser diode drive topology.

MOSFETs are usually overdriven to ensure that lowest available channel resistance is achieved, but a FET driven at a certain given voltage behaves like a current sink. This property can be exploited to provide the pulsed laser current. The idea is not applicable to the topology of Fig.4: a low  $R_{Ch}$  has to be used and then  $M$  cannot be used to set the laser diode current, because the major current flow will be through  $R_{Ch}$ . The topology shown in Fig.7 has been simulated using LT Spice,

with a gate voltage selected for every FET individually. Results are presented in Fig. 8.

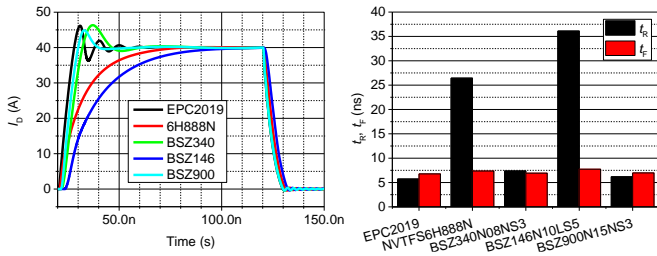


Fig. 8. FET current sink topology current waveforms (left) and speed (right).

Simulation indicates, that new idea can be used, however, some MOSFETs are too slow. Whilst this is expected for BSZ146N10LS5 (gate charge and reverse capacitance are 10 times larger than EPC2019), the results for NVTFS6H888N are quite unusual. The best rise times were for the EPC2019 (5.7 ns), the BSZ900N15NS3 (6.2 ns). The FET (EPC2019) current sink topology was selected as the candidate.

However, such a topology is not immune to temperature effects, and so additional modification can be used such as source feedback (Fig.9 left). This topology is usually used when a simple current sink is needed.

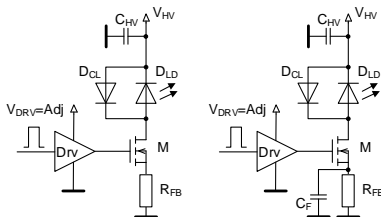


Fig. 9. FET with Source current feedback topology (left) and with filter capacitor to reduce parasitic L effects and boost the rising edge (right).

Feedback resistance  $R_{FB}$  can be chosen to provide few volts at maximum current. For a 40 A current and 100 mΩ  $R_{FB}$ , the feedback voltage would be 4 V. Such a low  $R_{FB}$  value becomes comparable to parasitic inductance: a 40 pH inductance can be expected, even if the flipped version is used [80]. It is not applicable to GaN FETs due to the high gate voltages used. A small filter capacitor  $C_F$  can be used in parallel to  $R_{FB}$  (Fig.9 right) to smooth parasitic L effects during signal transitions. This capacitor can also serve as a edge booster: FET current is increased if  $V_{FB}$  is lowered. These topologies were simulated using LT Spice with a 200 pH parasitic inductance for  $R_{FB}$ . Results are presented in Fig. 10 and Fig.11.

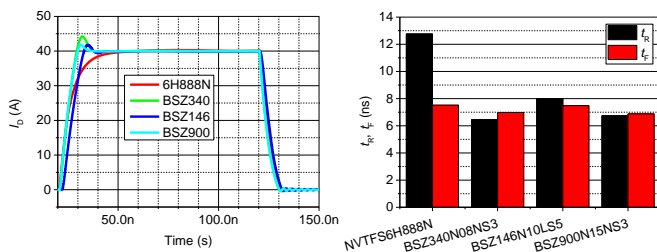


Fig. 10. FET source feedback current waveforms (left) and speed (right).

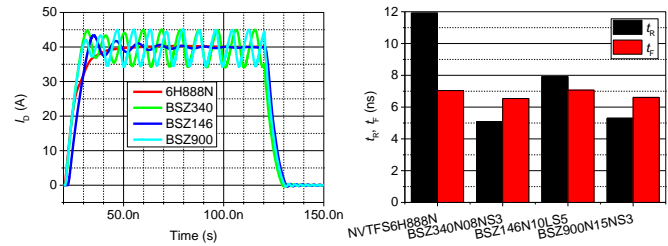


Fig. 11. Boosted FET source feedback waveforms (left) and speed (right).

The best rise times for the non-boosted versions was for BSZ340N08NS3 (6.5 ns) and BSZ900N15NS3 (6.7 ns). The boosted version (10 nF  $C_F$ ) introduced oscillations for BSZ340N08NS3 (5 ns) and BSZ900N15NS3 (5.3 ns), though the edges were the best for these FETs. Edge durations for NVTFS6H888N were not the best, yet it did not cause the oscillations thanks to the lowest reverse capacitance (note  $C_R$  value in Table II). The boosted FET (NVTFS6H888N) source feedback topology was selected as the candidate.

It can be concluded, that it is reasonable to expect rising and falling edges of 5-6 ns at 40 A pulsed current and a 40 V compliance voltage. It should be noted that the highest forward voltage laser diode was used in the simulation. Use of a lower forward voltage laser would give faster rise times.

Three topologies can satisfy the requirements for the generation of the APWP sets: a boosted resistor current limiter (Fig.1 right), a constant current sink FET (Fig.7) and a FET with source boost current feedback (Fig.9).

### E. Current Sensing

Current sensing is required in order to have feedback on the actual laser diode current. The most straightforward idea is to use a grounded current shunt resistor along a laser current path. While convenient for capacitor discharge (shunt resistor can be placed in series with  $D_{LD}$ , Fig. 4), such a solution is complicated for the constant current sink topologies: if a shunt resistor is placed in series with  $C_{HV}$ , then the charging current of this capacitor is also measured. Another solution would be to place a shunt at the high side and measure the differential voltage using a small signal transformer (Fig.12 left).

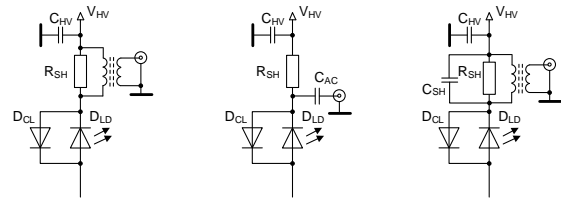


Fig. 12. High side current shunt using transformer (left), capacitor (center) and boosted shunt to reduce parasitic L effects (right) current sensing.

A standard small signal transformer would suffice, such as for example, the CX2041 from Pulse Engineering, Inc. It has dimensions of 3.8x3.8 mm, has a 200 μH magnetizing inductance  $L_m$ , and so lower cutoff frequency of 400 Hz with an  $R_{SH}$  of 500 mΩ. If  $R_{SH}$  is located near  $C_{HV}$ , there is no influence from parasitic inter-winding capacitance. It is also possible to replace the transformer by a capacitor (as per the middle circuit in Fig.12).

The main disadvantage with this sensing circuit is that low  $R_{SH}$  values are preferred, in order to keep the compliance voltage and efficiency high. In such a case, the parasitic inductance of  $R_{SH}$  might distort the measurement results by giving overshoots at edges. This is seen in the LT Spice simulation results on the left in Fig.13, for 50 m $\Omega$  and 0.6 nH parasitic inductance (three parallel standard wrap-around 0603 components) in Fig.13, for a FET current sink topology using EPC2019. Thanks to the low  $R_{SH}$  value, the current flowing through the laser did not change significantly (compare red and blue curves), yet the current sensing signal derived on the shunt has peaks and oscillations (black curve).

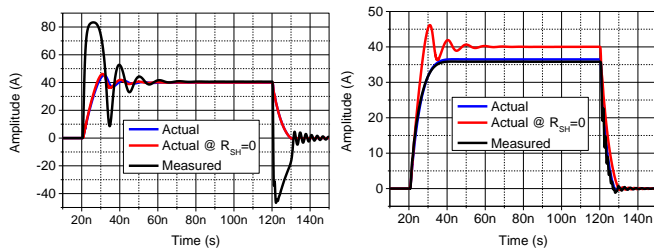


Fig. 13. Current sensing waveform of 50 m $\Omega$  shunt (left) and 500 m $\Omega$  shunt (right) comparison to actual current for FET current sink topology.

If the  $R_{SH}$  value is increased, e.g. up to 0.5  $\Omega$ , even switching oscillations disappear, but the rising edge time increases significantly, from 5 ns to 10 ns (Fig.13, right). Because of the higher  $R_{SH}$  value, the compliance voltage present on the laser is reduced by 20 V (0.5  $\Omega$ , 40 A). This causes not just the rise time reduction (refer eq. (1)), but also a decrease of the set current, since the current feedback has a limited gain (compare the red and blue curves: 36 A vs. 40 A). The resistor current limiter topology already has  $R_{SH}$  (actually it is  $R_{LIM}$ ) and its value is as required, 0.5  $\Omega$ , so that it can use this current sensing circuitry without any modifications. Acceptable results (Fig.14) can be achieved if  $R_{SH}$  is tampered by parallel capacitance (as per Fig.12 right) and  $R_{SH}$  is slightly increased to 100 m $\Omega$ .

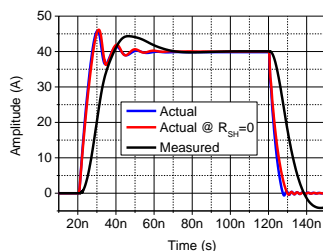


Fig. 14. Current sensing waveform of 100 m $\Omega$  shunt tampered by 100 nF: comparison to actual current for FET current sink topology using EPC2019.

Positive effect can also be noted on slightly improved edges (compare the blue and red curves). It should be noted that such a circuit corresponds to  $R_{FB}$  being used in a FET source feedback topology, so this topology also has inherent current sensing.

Another current sensing solution is a current transformer (Fig.15). If  $R_{SH}$  is low and the transformer turns ratio  $n$  is high then resistance appearing on the primary is lower by  $n^2$ . Even a small toroidal core can be used. In the case presented here, a

3.8 mm diameter toroidal core, taken from a CX2052 transformer (Pulse Engineering, Inc, San Diego, CA, USA) was used. It has 1.6  $\mu$ H per turn ( $A_L$  value), so a 1:10 turns ratio and 1  $\Omega$   $R_{SH}$  would result in a 1 kHz lower cutoff frequency, which is more than enough for the target APWP signals.

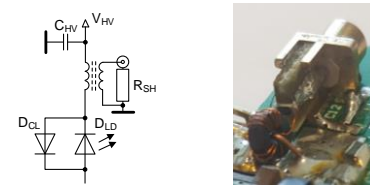


Fig. 15. High side current sensing using current transformer.

A low resistance, transformed into the primary would result in low voltages, which in turn will prevent the core from saturation. Unfortunately, a high turns ratio reduces the coupling  $k$ , so that some leakage inductance remains, which might reduce the rising and falling edge speed of the laser driver. The inductance of the secondary winding of the transformer was measured at 1 MHz, giving  $L_s=377$  nH (primary shorted) and  $L_2=168$   $\mu$ H (primary open) values. The coupling factor  $k$  was calculated as 0.999 according to [101]:

$$k = \sqrt{1 - \frac{L_s}{L_2}} \quad (4)$$

Such a coupling factor would mean that there is a 1.8 nH leakage inductance present on the primary, which is a significant fraction of the estimated 4.6 nH of total current loop inductance. This situation was simulated using LT Spice, and the results are presented in Fig.16.

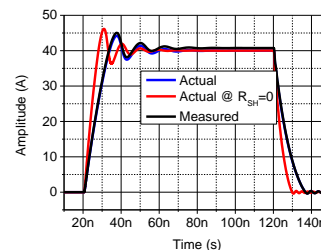


Fig. 16. Current sensing waveform when using 1:10 current transformer: comparison to actual current for FET current sink topology using EPC2019.

It is notable that the measured current (black curve) closely matches the actual laser current (blue curve), and rising and falling edges are reduced significantly: from 5 ns without current sensing transformer (red curve) to 10 ns with the transformer.

### F. Optical System

A simple optical system was used (Fig.17), comprising one aspheric 10 mm diameter, 8.13 mm focal length (FL at 905 nm) lens ACL108U-B (Thorlabs Inc., Newton, NJ, USA) placed at the 2FL (16.4mm) distance from the laser aperture.

Such a setup is compact and requires minimum adjustments and provides a magnification of 1:1, but at the expense of reduced beam coverage. For the largest fast axis beam divergence angle, 25 $^\circ$  (refer Table I), the beam full width half power (FWHP) size, when entering the lens, is 5.1 mm.



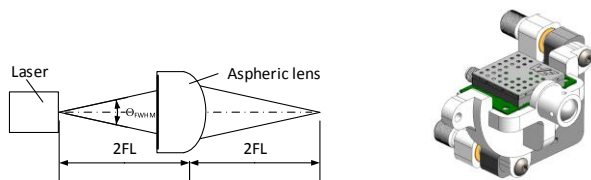


Fig. 17. Optical system schematic diagram (left) and assembled holder with kinematic mount 3D drawing (right).

The clear aperture of the lens (note thick line in Fig.17, which is to scale) is more than 9 mm, so that less than 0.2% power loss is expected through the optics.

#### IV. EXPERIMENTAL INVESTIGATION RESULTS

The output of the photodetector or preamplifier was digitized using a digital oscilloscope (Yokogawa DLM2054, 500 MHz bandwidth, equivalent to the 0.7 ns edge duration, 125 GS/s sampling rate with interpolation, Yokogawa Electric Corporation, Tokyo, Japan) when a single pulse or simple toneburst experiments were carried out. The laser system in these cases was driven by a Siglent SDG6022X arbitrary waveform generator (200 MHz bandwidth, 2.4 GS/s sampling rate, Siglent Technologies, Shenzhen, China).

##### A. Boosted MOSFET Source Current Feedback Laser Driver

The experimental investigation of the boosted Source current feedback topology (Fig.9, right) using the NVTFS6H888N MOSFET is analyzed here. The current loop inductance reduction was achieved from the PCB layout. The main current loop was routed on a 0.6 mm thick substrate (Fig.18).

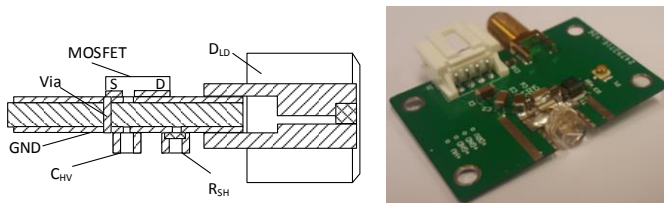


Fig. 18. Main current loop PCB cross-section (left) and assembled PCB (right).

The capacitor  $C_{HV}$  actually was constructed from a set of parallel connected 47  $\mu F$ , 4.7  $\mu F$ , and 0.47  $\mu F$  capacitors, with the smallest in the main loop (Fig.18) and the rest located away from the main loop. The distance from the MOSFET source to the PCB edge where the laser diode was soldered, was only 6 mm. The smallest capacitor  $C_{HV}$  and  $R_{SH}$  were of the inverse type 0612 size. The shunt resistor  $R_{SH}$  was a flipped type (resistive layer facing the PCB, note on Fig.18, left) and the laser diode  $D_{LD}$  had minimal pin length, so that the loop area was significantly reduced. The whole PCB was 39x27 mm in size, with the essential components occupying an area of 10x20 mm (Fig.18, right).

The current transformer better matches the laser current, as is seen in the comparison of current measured using current transformer (Fig.19, left) and  $R_{FB}$  (used as  $R_{SH}$ ) voltage (Fig.19,right).

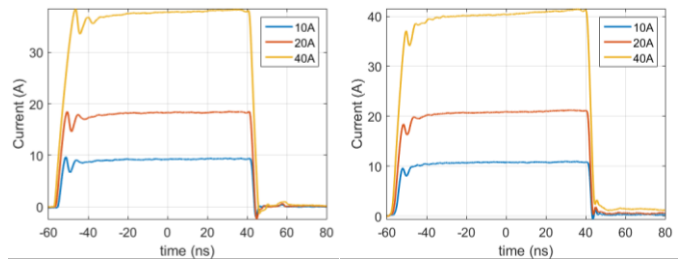


Fig. 19. Current transformer (left) and 100 mΩ shunt (right) current sensing.

The setup for the optical output waveforms measurement is presented in Fig.20. A fast (35 ps rise time) Si photodiode FDS015 (Thorlabs Inc., Newton, NJ, USA) was used, reverse biased with 5 V (Fig.20, left). The diode has an active area of 150  $\mu m$  diameter, which was further reduced by placing a 10  $\mu m$  diameter pinhole on top of the diode (Fig.20, right).



Fig. 20. Photodetector schematic diagram (left) and 3D drawing of the assembled sensor (right).

The current transformer has a slight detrimental effect on the switching speed (from 5.8 ns when no transformer is used to 7.2 ns with transformer at 40 A). Therefore, it was decided not to use current transformer for current sensing. All further experiments used the current shunt for laser current estimation. Laser driver current sensing was used to establish the required gate voltage and this voltage was used throughout succeeding experiments.

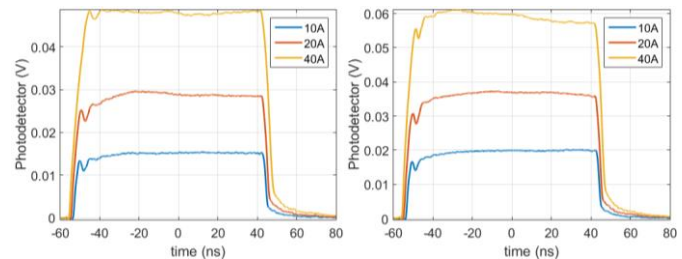


Fig. 21. Optical output with (left) and without (right) current transformer.

It is notable that there is some droop in optical output at 40 A. Explanation could be the laser efficiency reduction due to heating.

It should be noted that MOSFETs have larger reverse capacitance ( $C_{GD}=3$  pF, compare to 0.7 pF for GaN), therefore are sensitive to  $dV/dt$  effect when the fast falling drain voltage edge causes the MOSFET to turn off. Phenomenon can be mitigated by placing a small resistance in series with gate drive. Single channel gate driver and 3  $\Omega$  gate resistor was used to reduce  $dV/dt$  turn-off effects. The waveforms presented above are for the non-boosted case. See Fig.22 for the case when the filtering capacitor  $C_F$  is added in parallel to  $R_{FB}$ .

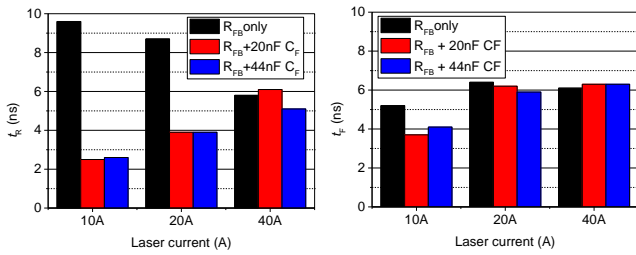


Fig. 22. Measured rising (left) and falling (right) edges for boosted source feedback driver vs. boost capacitance.

Pulses with widths down to 20 ns can be produced using the 20 nF boosted case (see Fig.23) for a 20 ns pulse train with a 40 ns period. Notable that current amplitude of pulses is the same for every programmed current (Fig.23, right). Amplitude of pulses is dropping slightly (4 %) at 40 A current. This can be explained by laser efficiency reduction due to heating.

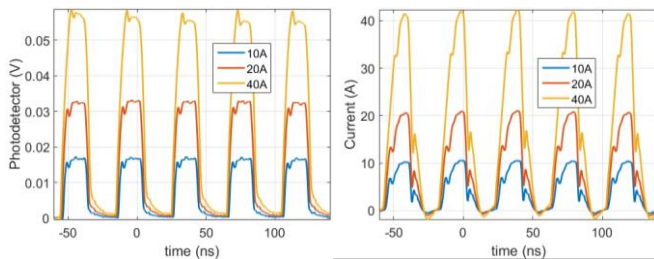


Fig. 23. Pulse train optical output (left) and current (right) for boosted source feedback topology.

It can be concluded that down to 20 ns pulse sets can be generated using NVTFS6H888N MOSFET in boosted source current feedback topology.

### B. GaN Boosted Resistor Current Limiter Laser Driver

Experimental investigation of the boosted resistor current limiter (Fig.1, right), using EPC2019 GaN FET driven by LMG1020 is analyzed in this section. The same approach as shown in Fig.18 was used to minimize the current loop inductance, and the PCB was made from a 0.6 mm substrate. The current limiting resistance  $R_{LIM}$  was 500 m $\Omega$ , and the capacitance used for boosting the edge current,  $C_{BOOST}$  was 8 nF (which was located in the main loop as  $C_{HV}$  in Fig.18). The same limiting resistor  $R_{LIM}$  was used as a current shunt (see the waveform obtained in Fig.24).

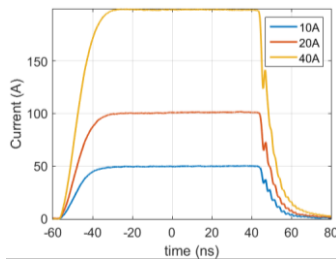


Fig. 24. Current sensing waveform for 500 m $\Omega$  shunt in case of boosted resistor current limiter topology laser driver.

The transition time looks slow because of the smoothing effect of the boosting capacitance, but the actual laser current so the optical output edges are steeper (refer Fig.25).

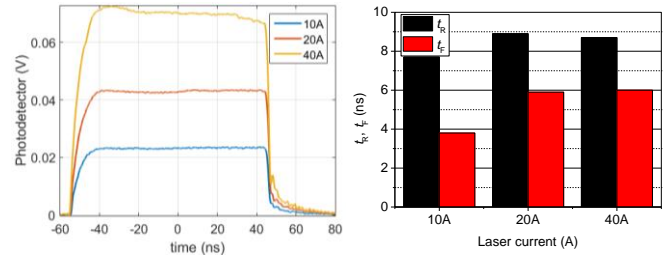


Fig. 25. Optical output waveform (left) and speed (right) for boosted resistor current limiter driver.

The rise time is slower at low currents because of lower compliance voltage. Voltage  $V_{HV}$  had to be varied to get the corresponding laser current (33 V for 40 A, 20 V for 20 A and 12 V for 10 A).

Pulses as short as 20 ns can be produced, as is seen for the pulse train with 20 ns pulses with a 40 ns period in Fig.26.

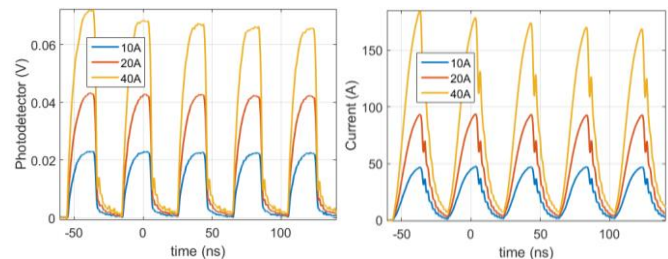


Fig. 26. Pulse train optical output (left) and current (right) for boosted resistor current limiter driver.

It can be noted, that pulse amplitude is stable, only 40 A current has a 9 % drop between last and first pulse. It is more pronounced than boosted source feedback (4 %) because the laser heating and the current drop combine.

### C. GaN Constant Current Sink Laser Driver

The experimental investigation of the constant current sink FET (Fig.7) using the EPC2019 GaN FET is analyzed in this section. The same approach as for Fig.18 was used to minimize the current loop inductance, and the PCB was made on 0.6 mm substrate. The current was measured using a 100 m $\Omega$   $R_{SH}$  as per Fig.12, (right). The shunt resistance was tampered by the 20 nF capacitor, to filter the oscillations caused by the parasitic inductance, and to provide some boost on the edges. The waveform recorded on  $R_{SH}$  is presented in Fig.27.

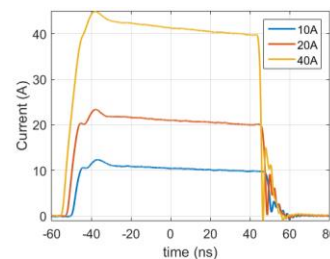


Fig. 27. Current sensing waveform for 100 m $\Omega$  shunt for current sink driver.

Note that there is some peaking in the current, the same is registered on the optical output (Fig.28, left). Measured edges of the optical output are presented in Fig. 28, right.

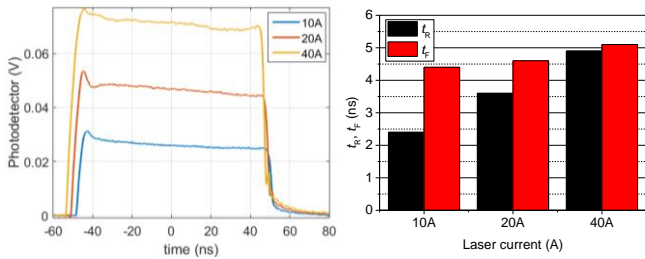


Fig. 28. Optical output waveform (left) and speed (right) for constant current sink FET laser driver.

Note also, that there is a drop in the optical output and laser current (Fig.27). The possible reason could be the heating of the constant current FET: all of the compliance voltage drop is on the FET, so that significant instant power is dissipated on the FET. The drain voltage according to Fig.29 is 24 V, which corresponds to 960 W instantaneous power.

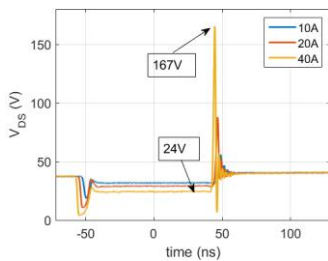


Fig. 29.  $V_{DS}$  voltage for constant current sink FET laser driver.

The clamping diode  $D_{CL}$  was placed close to the laser, to reduce the falling edge duration, the penalty for doing so being a 167 V drain voltage induced after FET turn off (which is still below the 200 V breakdown according to Table I).

The waveform of the optical output in response to the pulse train (20 ns duration, 40 ns period) is presented in Fig.30.

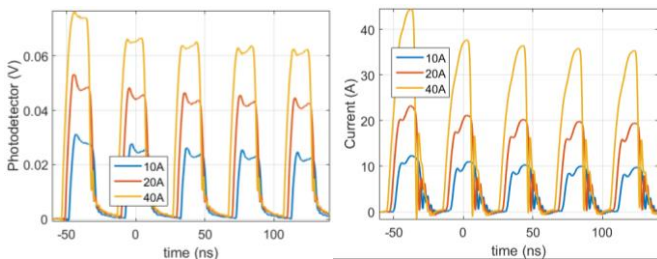


Fig. 30. Pulse train optical output (left) and current (right) for constant current sink FET laser driver.

The pulse amplitude drop for this topology is the most visible, with the last pulse amplitude (62 mV) being 19% less than the first pulse amplitude (76 mV). The drop is present for both high and low driving currents.

It is possible to program the rising edge duration by changing the  $V_{HV}$  voltage, and adjusting the gate drive voltage. See the signals obtained on the Fig.31, left (gate voltage was adjusted to maintain a current of 40 A). The attainable rising edge duration versus the compliance voltage is presented in Fig.31, right. The thin lines are theoretical estimates, derived from (1):

$$t_R = \frac{I_{LD} \cdot L}{V_{HV} - V_{LD}} + t_{drv}, \quad (5)$$

where  $I_{LD}$  is the laser diode current,  $V_{LD}$  is the laser forward voltage (can be obtained from Fig.29),  $t_{drv}$  is the additional delay from gate driver, and  $L$  is the main current loop inductance. The main current loop inductance (3.2 nH) was estimated by fitting (5) to Fig.31 results.

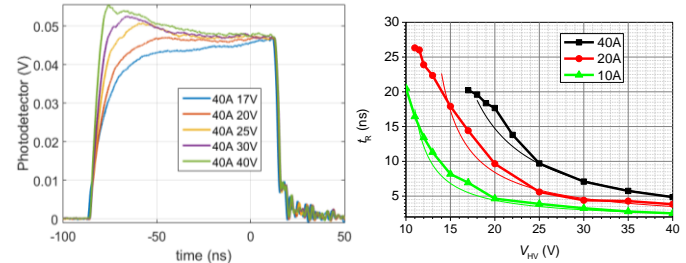


Fig. 31. Optical output variation with different compliance voltages (left) and the attainable rising edge duration (right) versus compliance voltage for constant current sink FET laser driver.

The GaN constant current sink FET topology was used in further measurements.

### D. Beam Profile

Beam profile was measured using the 1280x1024 pixels monochromatic CMOS camera [102] and 1:4 beam expander. A pixel pitch of  $5.07 \mu\text{m}$  was estimated using an optical calibration scale. The beam profile images obtained (Fig.32, Fig.34) were used for beam width measurements. A slice of the measured profile (black curves) along the x axis (parallel to the junction, slow axis, Fig.33 and Fig.35, left) or along the y axis (fast axis, Fig.33, Fig.35, right) was approximated (blue curves) by a Gaussian function with a DC offset [103] in order to extract the full width half power (FWHP) beam size.

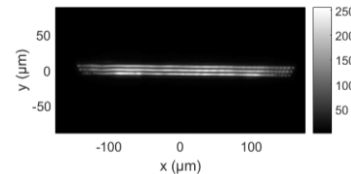


Fig. 32. 2D near-field beam profile.

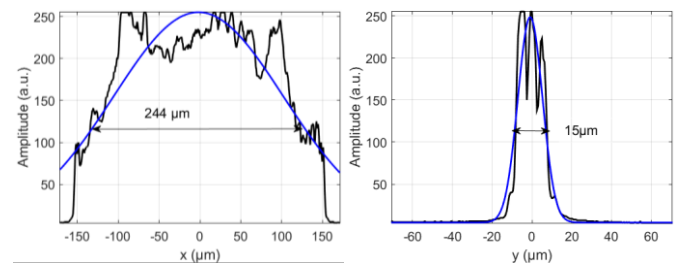


Fig. 33. Measured (black curve) and approximated (blue curve) near-field beam profile along x axis (left) and y axis (right).

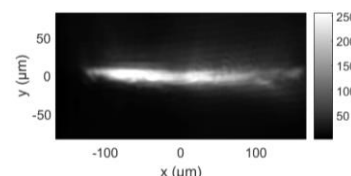


Fig. 34. 2D focused laser beam profile.

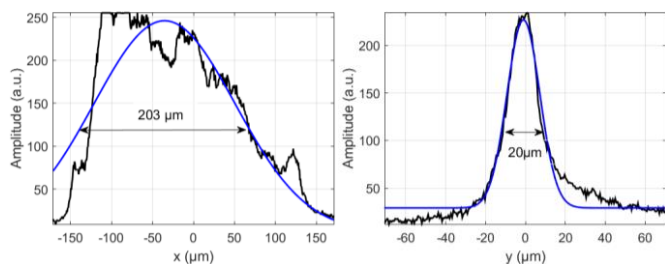


Fig. 35. Focused laser beam profile along x axis (left) and y axis (right): measured (black) and approximated (blue).

It was found that the beam spot was subject to comma and other off-axis defects, since there was no laser beam nor lens alignment. This can be seen by comparing the near-field beam profile (Fig. 32, Fig.33) to focal spot images in Fig.34, Fig.35.

It should be noted that asymmetric beam profile might influence the produced acoustic output directivity: such a beam profile is equivalent to line source excitation [22],[74],[104]-[106]. If the surface is not constrained, the Rayleigh surface waves will dominate. This property can be attractive in applications where surface defects have to be detected [16],[35],[105]. In this paper, the target is the longitudinal wave generation (propagating normal to the surface). Fortunately, in the case of the constrained acoustic source, the amplitude of the longitudinal waves is significantly enhanced [44].

### E. Acoustic Output

The photoacoustic response was measured using the ultrasonic transducer with its receiving surface exposed to the focused laser beam (Fig.36). The surface of the transducer was coated by an absorbing coating (black permanent marker, Stanley, Fine Tip). The transducer was screwed into a 4 mm thick aluminum plate, which in turn was bolted to a 10 mm thick clear polycarbonate (PC) disk. The PC disk served as a constraining layer to increase the longitudinal wave amplitude directed normal to the surface [38]-[45].

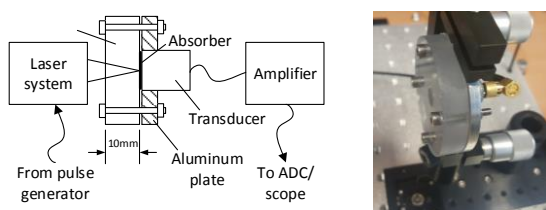


Fig. 36. Setup for photoacoustic response measurement diagram (left) and photo of the assembled clear polycarbonate (PC) disk with transducer (right).

Three thread-in, contact type 6 mm active element diameter transducers were used: a 2 MHz C542-SM, a 5 MHz C543-SM and a 10 MHz C544-SM (Olympus Corp., Westborough, MA, USA). The transducers were of the CENTRISCAN design, which offers both high sensitivity and wide bandwidth. The transducer output was fed into the wideband (0.5 MHz to 30 MHz range) programmable gain (7 dB to 47 dB range) preamplifier [107].

An example of the acoustic output for the 5 MHz ultrasonic transducer for different pulse duration (1, 2 and 4 periods) and amplitude (10 A for 4 periods to 40 A for one pulse) toneburst signal and its spectrum are presented in Fig.37.

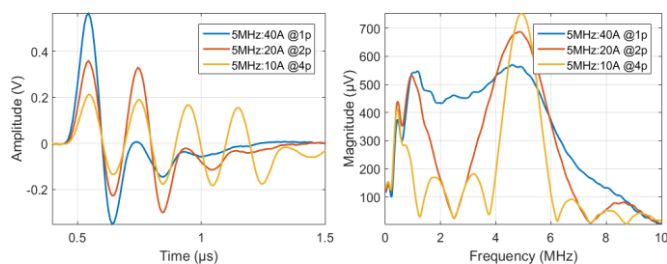


Fig. 37. Photoacoustic response in time (left) and frequency (right) domain for 5 MHz transducer.

The response for 2 MHz and 10 MHz transducer when a 10 mm PEEK (polyetheretherketone) was inserted between polycarbonate and transducer is presented in Fig.38.

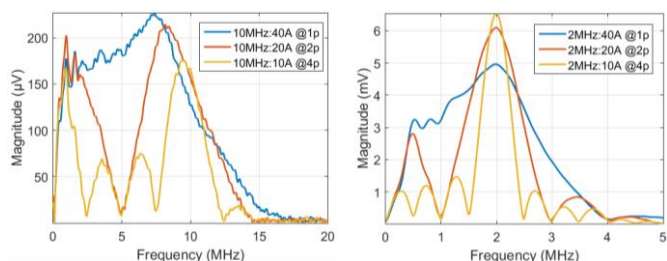


Fig. 38. Photoacoustic response spectrum for 10 MHz (left) and 2 MHz (right) transducer after propagation through 10 mm PEEK.

The attenuation effects can be seen on Fig.38, left: the high frequency components have been attenuated while propagating in plastic. The experiments reported above demonstrate that the signal frequency content can be concentrated in the band of interest, when using a pulse sequence. Meanwhile, a single pulse has a lot of energy concentrated close to DC, which is of little use for ultrasonic measurements.

A dedicated ultrasonic signals acquisition system was used both to drive the laser system (binary code sets at 100 MHz sampling rate) and to digitize the preamplifier output (10 bit, 100 MHz sampling rate) [107], when generation of the complex APWP sequences was required.

A demonstration of the more complex APWP spread spectrum [108],[109] waveforms (at 10 A current, when using the SPL-PL90-3 laser) for a 5 MHz transducer surface probing is presented in Fig.39 (one period chip PSK signal coded by 13 elements Barker code) and Fig.40 (6.5 μs duration chirp with 10 MHz to 3 MHz frequency sweep).

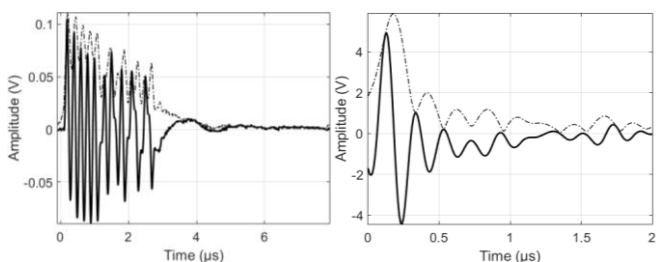


Fig. 39. Raw (left) and compressed (right) output for 5 MHz transducer excited by PSK signal coded by 13 elements Barker code (dashed line - envelope).

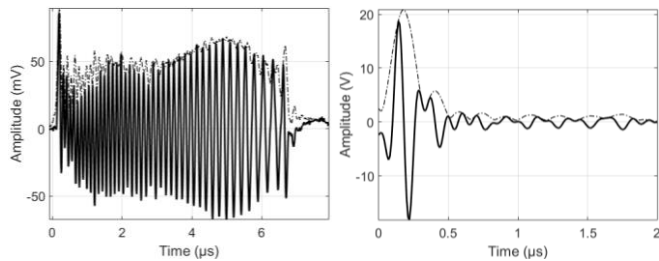


Fig. 40. Raw (left) and compressed (right) output for 5 MHz transducer excited by chirp with 10 MHz to 3 MHz frequency sweep (dashed line – envelope).

It can be seen that amplitude of the signal is increased (compared to a single pulse response in Fig.41).

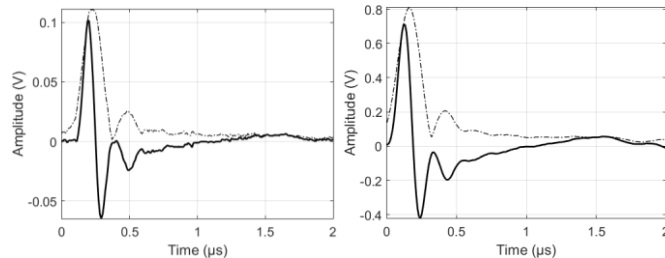


Fig. 41. Raw (left) and compressed (right) output for 5 MHz transducer excited by a single pulse (dashed line – envelope).

The experiments above have been carried out at current of 10 A, because safe operation area of the laser does not permit one to use such long pulse sets. Higher currents can be used with the new generations of laser diode that are now available.

## V. CONCLUSION

A compact driver for laser ultrasonics experiments with arbitrary position and width pulse sequences has been developed. While it was concluded that GaN based constant current switch topology has best performance, power MOSFET in source current feedback topology has a comparable performance. Latter topology produces more stable amplitude. Both topologies can supply up to 40 A pulse sets, where individual pulse duration can vary between 20 ns to 1000 ns, which allows one to cover frequencies in the 0.5-25 MHz ultrasound range. Shorter pulse durations can be attained, but at durations below 10 ns the pulse amplitude decreases. Longer pulse durations beyond 1000 ns can be used, but there is some output decrease beyond this point due to FET and laser heating. It has been demonstrated that the acoustic outputs for 2 MHz, 5 MHz and 10 MHz transducers can be programmed, by changing the spectral content of the excitation signal via complex spread spectrum signals (PSK or chirp). The further development could be concentrated on the photoacoustic output and directivity, in relation to the optical system and excitation set design studies.

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