Fast high-voltage amplifiers for driving electro-optic modulators

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We describe five high-voltage (60 to 550 V peak to peak), high-speed (1-300 ns rise time; 1.3-300 MHz bandwidth) linear amplifiers for driving capacitive or resistive loads such as electro-optic modulators. The amplifiers use bipolar transistors in various topologies. Two use electron tubes to overcome the speed limitations of high-voltage semiconductors. All amplifiers have been built. Measured performance data is given for each.

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I. INTRODUCTION

Electro–optic modulators (EOMs) are used for amplitude, frequency, and phase modulation of laser beams and, for example, as actuators in optical frequency and phase locked-loops. They typically require input voltages in the range of \( U_r = 50 - 500 \) V for generating an optical phase-shift of \( \pi \). Because fast amplifiers capable of driving modulators with these voltages are difficult to make, the overall speed is usually limited by the driver amplifier.

The best (monolithic or hybride) operational amplifiers achieve a slew rate \( \frac{dU}{dt} \approx 5000 \text{ V/µs} \), but only at low voltages of \( \pm 15 \) V. At high voltage, hybride op amps achieve \( \frac{dU}{dt} \approx 1000 \text{ V/µs} \) at 400 Vpp (volts peak to peak) output amplitude, i.e., rise-times of 400 µs and a large signal bandwidth of 0.8 MHz. Amplifiers based on discrete MOSFETs achieve similar performance. For example, amplifiers made by various companies provide up to 800 Vpp for sinewaves at frequencies up to 250 kHz, i.e., \( \frac{dU}{dt} = 600 \text{ V/µs} \), and twice the voltages in a push-pull configuration, where both electrodes are driven by paraphase signals to double the voltage. Integrated amplifiers for driving CRTs in high-resolution monitors achieve about 600 Vpp output voltage and 3 ns rise-time for low capacity loads (< 20 pF). EOMs can also be driven by RF wide-band amplifiers, preferably in push-pull configuration, but the output voltage of these is generally limited to below 100 Vpp.

In applications like frequency and phase locked loops for diode-, Titanium:sapphire-, or dye- lasers, speed and voltage demands can barely be fulfilled by existing amplifiers; such setups can thus be improved (or simplified) by an improved high-voltage amplifier. Electrically, an EOM represents a capacitive load of the order of 10-150 pF. There is a tradeoff between high speed and high output voltage: because the current required to charge the EOM and circuit capacity increases with frequency and voltage, the power dissipation of amplifiers based on a given technology is roughly proportional to the bandwidth times the square of the maximum output voltage; moreover, high voltage semiconductors tend to be slower. In this article, we describe five amplifiers that can drive capacitive loads with different combinations of high speed and output voltage. They use several tricks to achieve high performance at moderate power dissipation. In Sec. A we present a push-pull bipolar transistor design which achieves 550 Vpp output voltage and > 1.3 MHz bandwidth, about twice the one of high-voltage operational amplifiers. Section B describes an amplifier whose output voltage and bandwidth of 65 Vpp and 75 MHz are comparable to those of high-resolution CRT drivers, but that can drive loads of up to 100 pF. Section C describes a nanosecond rise-time push-pull amplifier for 60 Vpp. At high output voltages, one can achieve improved performance by using electron tubes: Section D describes an amplifier for a maximum output voltage of 550 Vpp and > 5.7 MHz bandwidth, more than four times faster than our transistorized amplifier for 550 V. In Sec. E we describe a 140 Vpp amplifier for > 120 MHz bandwidth.

II. BIPOLAR TRANSISTOR DESIGNS

While MOSFETs are intrinsically very fast, their high input and output capacity are significant disadvantages for their use as fast linear amplifiers. For example, an amplifier capable of generating 400 Vpp output voltage with 300 V/µs slew rate is described in Ref. [6], intended for the use as a piezo driver. It uses a MTP2P50E (p-channel) and an IRF830 (n-channel) MOSFET, whose output capacities add to 220 pF. While this is much lower than the capacity of a typical piezo, it is greater than the capacity of a typical EOM (100 pF). Thus, when driving an EOM, a large part of the amplifier’s potential output current is used to charge the internal capacities. This reduces the potential speed of MOSFET amplifiers. Bipolar transistors, on the other hand, are available that have < 3 pF output capacity.

A. 550 V pp, 300 ns differential amplifier

This amplifier (Fig. 1) uses a differential amplifier configuration with complementary emitter followers for pro-
FIG. 1: Schematic of the 550Vpp, 300ns amplifier.

viding a 550Vpp output voltage with relatively low power dissipation. A positive and a negative signal each provide half the output signal amplitude. With zero input signal, the output voltage across the load is zero. This amplifier can be used for any loads that allow for a differential drive, such as EOMs made by LINOS photonics.

The main building block is the differential amplifier consisting of T7 and T10. These transistors are connected with T3 and T4, respectively, to form two cascode stages. T7 and T10 act as current amplifiers. They can be low-voltage types BFW16A that feature a transition frequency $f_t > 1$ GHz. Voltage amplification is done by T3 and T4, which must be high voltage types. Since they operate in a common-base configuration, their transition frequency $f_t$ can be considerably lower without degrading the overall speed. Moreover, they can be used up to their $U_{CEO}$ voltage rating, which usually is 20 to 50% higher than $U_{CEO}$, the maximum value for common-emitter configuration. We choose types BF759 and BF761, that feature $U_{CEO} = 350$ V, a maximum total power dissipation $P_{tot} = 10$ W, $f_t = 45$ MHz and about 3.5 pF output capacity. (These types seem to be out of production, though they are still available in small quantities. Tested substitutes that even lead to slightly improved speed due to their lower capacities are 2SB1011 (pnp) and 2SC3063 (npn).) T8 and T9 operate as constant current sources for the differential amplifier. The output signals are delivered to the load via two complementary emitter followers, one consisting of T1 and T5, the other of T2 and T6. They are biased for class AB operation by diodes 1N4148 to reduce the crossover distortion.

The gain is set to 100 by $(R_{28} + R_{29})/R$, where $R$ is the resistance of the network consisting of R16, R17, and R21. The bandwidth of the amplifier is determined mainly by the low-pass filter formed by $R_{28} = R_{29}$ and the transistor and wiring capacity at the node at the collectors of T3 or T4, respectively, that amounts to about 15 pF. This gives a theoretical -3dB–bandwidth $B = 1/(2\pi R C) \approx 1.1$ MHz. The inductors L1 and L2 increase the effective load impedance at high frequencies, which leads to a theoretical increase of the bandwidth by $\sim 40\%$ with 1% overshoot. An additional increase of the bandwidth is provided by C5 and C16. The inductors and capacitors are adjusted for optimum square-wave response at a high signal voltage (500Vpp).

R28 and R29 dissipate about 2.25 W each at zero input signal. However, they must be rated for 10 W so that the amplifier can be continuously operated with a DC signal of full amplitude. T3 and T4 dissipate at most 2.25 W each under all signal conditions; a small heat sink (mounted with minimum stray capacity) is attached to them. Each of T1, T2, T5, and T6 will dissipate 3.75 W at full output voltage with a 500 kHz sinewave and 100 pF load; cooling was provided by attaching the transistors to the metal housing with an insulating layer.

The amplifier was tested with an LM0202 EOM (LINOS AG) as a load, that is specified to have 82 pF capacity, connected via short (25 cm) RG-58 cables (the use of low-capacity types such as RG-63 should be preferred for longer cables). The results are summarized in table II.

B. Single-ended 65 Vpp, 5ns amplifier with multiple output stages

EOMs made from materials such as lithium tantalate (LTA) can have $U_{\pi} < 50$ V. Such voltages can, in principle, be provided by integrated circuits intended for driving cathode-ray tubes in high-resolution monitors. However, while these work well with a low-capacity load of around 10 pF, they cannot provide the output current of 1.2 A peak that is required to drive a 100 pF load with 60 Vpp at 5 ns rise time. The amplifier presented in Fig. 2 delivers up to 85 Vpp into a single ended 100 pF load at less than 5 ns rise time.
FIG. 2: Single-ended 65 Vpp, 5ns amplifier with multiple complementary emitter follower output stage. C7-10 represent the load.

**TABLE I:** Technical data of the 550 V differential amplifier with a 100 pF load. With lower capacity loads, speed is up to 50% larger. Phase shift is about 45° at the -3dB frequency limit.

| Parameter                  | Condition | Value     |
|----------------------------|-----------|-----------|
| Gain                       |           | 100       |
| Input impedance            |           | 50Ω<sup>a</sup> |
| Max. output amplitude      |           | 550 Vpp   |
| Supply                     | +300V/80mA| -15V/30mA |
|                            |           | 10V/20mA  |
| Bandwidth                  | \( U_2 \leq 200 \text{ Vpp} \) | 2.0MHz   |
|                            | \( U_2 = 500 \text{ Vpp} \)   | 1.3MHz   |
| Rise time                  | \( U_2 \leq 200 \text{ Vpp} \) | 150ns    |
|                            | \( U_2 = 500 \text{ Vpp} \)   | 300ns    |
| Delay                      | \( U_2 \leq 200 \text{ Vpp} \) | 80ns     |
|                            | \( U_2 = 500 \text{ Vpp} \)   | 180ns    |

<sup>a</sup>determined by R1

Since a single output stage cannot provide the necessary current, we use four complementary emitter output stages in parallel. Each can drive an individual load of 25 pF. Combining their outputs, a single 100 pF load can be driven. Without the emitter followers, the amplifier can drive a single 25 pF load connected directly to the emitter of T3. The output transistors BFQ 262 and BFQ 252, made for the video output stages of high-resolution monitors, have a \( V_{cb0} \) of 100 V (115 V for BFQ252A/BFQ262A), \( f_T > 1 \text{GHz} \) and collector-base capacities \( C_{cb} \) of 2.5 pF (BFQ252) and 2 pF (BFQ262), respectively. They feature a 5 W power dissipation. The emitter followers operate without bias in class B, which causes some crossover distortion. When low distortion is important, they can be biased for class AB as the output emitter followers in Fig. 1.

The emitter followers are driven by a voltage amplifier output stage consisting of T1, T2, and T3. T1 and T2 are arranged in a cascode configuration, with the advantages discussed in Sec. IIA. The load is connected via the emitter follower T3. Since T3 can only source but not sink current, with a capacitive load (such as the complementary emitter follower output stages) the rise time for negative voltage changes would be much longer than for positive ones. Thus, the Schottky diode D2 has been connected from the emitter of T3 to the collector of T2. For positive voltage changes, D2 is reverse biased, but for negative ones, T2 can sink current from the load through D2. Thereby, negative voltage changes can be as fast as positive without an additional transistor. Alternatively, a complementary emitter follower could be connected to T2’s collector, but then the collector-base capacities of three (rather than two) transistors would contribute to the capacity at T2’s collector, thus reducing the speed. Circuits of this type are also used for driving cathode-ray tubes in high-resolution monitors.

Adjustable DC bias (nominally 2.3 V) is applied by T8 at the emitter of T1. Negative feedback via R9 and R10 sets the gain and gives the stage a low output impedance. A common emitter amplifier Q2 drives the stage. The network at Q2’s emitter compensates for the gain loss due to stray capacities parallel to R9 and R10 at high frequencies. The input signal is terminated into 50 Ω (R4). A pnp emitter follower Q1 drives the common emitter amplifier Q2; D1 shifts the DC level of the input voltage.

R11 must be mounted with low stray capacity. Since the negative feedback reduces waveform distortions, such as ringing, it may be a wire-wound type. Actually, the inductance of a wire-wound resistor even improves the risetime of this amplifier by series peaking. Under quiescent conditions, T2 dissipates at most 3.5 W. At 50 MHz, 50 Vpp and 25 pF load per output, each of the emitter follower output transistors dissipates 2.5 W. The amplifier was constructed with all BFQ252 and BFQ262 transis-
TABLE II: Technical data of the 65 Vpp amplifier.

| Parameter              | Value                        |
|------------------------|------------------------------|
| Output voltage range   | 10-75V<sup>a</sup>          |
| Input voltage range    | 0-1V                         |
| Gain                   | 60                           |
| Input impedance        | 50Ω                          |
| Load                   | 4 × 25pF or 1 × 100pF       |
| Rise time (60 Vpp)     | 5ns                          |
| Bandwidth              | 75MHz                        |
| slew rate              | 3 kV/μs                      |
| Supply<sup>b</sup>     | +80V/200mA; +5V/10mA        |

<sup>a</sup>Can be increased to at most 10-95V by rising the supply voltage to 100V

<sup>b</sup>at 50Vpp, 50MHz and 4 × 25pF load

C. 60 Vpp nanosecond rise-time amplifier with electronic amplitude control

The amplifier shown in Fig. 3 is designed for amplifying ECL level pulses, providing 60 Vpp pulses with nanosecond rise time into a symmetric low capacity load. The output amplitude can be adjusted continuously from zero to 60 Vpp by a DC voltage.

The input stage is designed for amplitude control. It is a differential amplifier (Q1 and Q2) with an adjustable constant-current source (Q4). The output amplitude is adjustable with a voltage at the base of Q3 between -5 V and -4 V. The collector capacity of the BFR182 transistors is only a few tenths of pF, so R5 can have a relatively high value of 100Ω, provided that wiring capacities are minimized.

An emitter follower Q5 provides a low impedance signal to the subsequent stage, a differential amplifier consisting of Q8 and Q9. It has threefold voltage gain. Its load resistance must be as low as 33Ω, since the more powerful transistors have higher capacities. This stage derives an adjustable supply voltage from Q7. Emitter followers T1 and T2 using the relatively powerful BFG235 transistor drive the power stage. Because of the low impedances here and in the power stage, the layout must be designed for low parasitic inductances. The lead lengths to the bases of T16 and T17 have to be made as short as possible, in order to reach the bandwidth of above 300 MHz required for the short rise-time. The BFG235s dissipate about 1W through their collector connections.

The output capacity of the power transistors must be as low as 6.6 pF to achieve the desired speed. They also have to meet a demanding combination of voltage, current, dissipation, and <i>f</i><sub>T</sub> ratings. Only UHF power transistors satisfy these. Many devices, however, come with a relatively narrow band internal impedance matching network and are thus not usable for this application. Transistors BLW33 have been chosen.

To eliminate the Miller effect (which would result in an unrealistically small input impedance), a cascode configuration is the only choice. This has the additional advantage that the “upper” transistor can be utilized up to <i>U</i><sub>cb</sub><sup>0</sup>, which is crucial because of the relatively low maximum voltages of UHF transistors.

The power stage uses negative feedback by emitter resistors. Each consists of five paralleled resistors (12Ω each) in order to minimize the parasitic inductance to ≪1nH. Also the resistors R29, 30, 31, and 42 must be very low inductance types; suitable are those that come in power transistor packages that can be mounted to a heat sink. L1 and L2 compensate for the gain loss at high frequencies; they consist entirely of resistor and wiring inductances. The power stage dissipates 27 W. The power resistors and transistors were mounted to heat sink. A fan was used to keep the heat sink below 50°C.

The technical data summarized in Tab. III refer to a circuit driven from a high-speed ECL waveform with a rise time < 0.3 ns. The load can be connected through two 50Ω coax cables of equal length. They are terminated inside the amplifier, so termination at the load is unnecessary (and would reduce the output voltage to one half).

III. ELECTRON-TUBE DESIGNS

Electron tubes are well suited for high-voltage linear amplifiers. At a given output capacity, they pro-
provide higher power dissipation and current capabilities than transistors (bipolar or MOSFET).

A. Single-ended 550 Vpp, 60 ns rise time amplifier

The power stage of the amplifier shown in Fig. 4 uses two tubes: V1 to source current to the load and V2 to sink current. The power stage is controlled by the control grid g1 of V2. Under quiescent conditions, this is at about $-23\,\text{V}$ relative to its cathode. Its corresponding anode current of $I_{a,0} = 85\,\text{mA}$ causes a voltage drop across $R_2$ that causes a similar negative voltage of $-23\,\text{V}$ at the control grid of V1 with respect to its cathode. If V2 is driven to conduct lower current, there will be less voltage drop across $R_2$, causing V1 to conduct more current. If, on the other hand, V2 is conducting a large current, V1 will be cut off by the voltage drop across $R_2$. Thus, the two tubes act as a push-pull amplifier. The circuit can therefore both source and sink peak currents $I_{a,m}$ of more than $800\,\text{mA}$, substantially larger than the quiescent current $I_{a,0}$ [11]. This makes a high slew rate $dU/dt$ possible at moderate power dissipation.

A differential amplifier (T5 and T6) is used as driver stage. The driver is significantly faster than the power stage. It uses the BFQ252 and BFQ262 described in section [11B]. The power stage is driven by a complementary emitter follower (T3 and T4).

Negative feedback over the power stage by $R_{10}$ is used to set the gain and provide a low output impedance. The output stage without feedback acts like an integrator (providing constant output current to a capacitive load). Since the feedback is over the power stage and the much faster emitter follower, the amplifier does not get unstable even with large capacitive loads. A small capacitor in parallel to $R_{10}$ stabilizes the amplifier. It is effectuated by attaching a thin copper strap on the surface of $R_{10}$, that is connected to the output side of the resistor. The overall gain of the amplifier is set by $R_{10}/R_{11}$. The high frequency compensation network consisting of $C_9, C_{12}, C_{13}$, and $R_{21}$ compensates for the distributed feedback capacity parallel to $R_{10}$. $R_{23}$ and $L_3$ stabilize the amplifier for capacitive loads.

a. The g3-trick. The slew rate of the amplifier without load $I_{a,0}/C$ is given by the capacity $C = C_2 + 2C_{ag1} + C_{stray}$ at the anode of V2, that consists of the output capacity $C_2$ of V2, the anode-to-g1 capacity $C_{ag1} = 2.5\,\text{pF}$ of V1 and V2, and stray capacities $C_{stray}$. The $C_2$ of a PL509 pentode is specified as $17\,\text{pF}$ with g3 connected to cathode [12, 13], so $C \approx 25\,\text{pF}$ if $C_{stray} \approx 3\,\text{pF}$. The output capacity $C_2$ of V2 can be reduced to $9\,\text{pF}$ by the 'g3-trick', i.e., by connecting the g3 of this tube to the cathode via $R_3$ rather than directly, which allows the beam plate g3 to have a floating RF potential. $C_2$ can be further reduced by connecting this g3 to the output (preferably through a $100\,\text{V}$ Zener bypassed with a $10\,\text{nF}$ capacitor to make g3 negative with respect to the anode), thereby practically eliminating the g3-to-anode contribu-
tion to the capacity. From the measured rise-time (Tab. IV) and \( I_{a,0} = 85 \text{mA} \), \( C \approx 12 \text{pF} \) can be calculated, lower by a factor of over 2 compared to the above value, i.e, \( C_2 \) has been reduced fourfold to about \( 4 \text{pF} \).

b. The \( L_k \) trick The maximum slew rate of the capacitively loaded amplifier is proportional to the peak anode current \( I_{a,m} \), which charges the load and circuit capacities. However, without \( L_2 \), the slew rate for positive voltage steps at the output would be significantly lower, since the current required to charge the anode capacity of \( V_2 \) is provided by \( V_1 \) via \( R_2 \). This current causes a voltage drop across \( R_2 \), which makes the grid of \( V_1 \) negative. This reduces the anode peak current that is available from \( V_1 \) and leads to a reduced speed of positive transitions. With \( L_2 \), however, a sudden change in the anode current of \( V_2 \) cause a positive voltage spike at the grid of \( V_1 \). In the limit of a very large \( L_2 \), a small negative change of \( V_2 \)’s anode current will drive \( V_1 \) fully open (in fact, the grid voltage of the upper tube may even get positive). The slew rate for positive edges is thus increased. \( L_2 \) is bypassed for negative edges by \( D_4 \) and the 27V Zeners \( D_3 \) and \( D_4 \). The Zener is to allow a \( \leq 27V \) drop across \( L_2 \) for negative slopes, which is necessary because the arithmetic mean of the voltage across an inductance is always zero.

c. Power supplies and dissipation A Voltage-doubler circuit (TR3,D10,D11,C15,C16) provides the positive DC voltages. A negative supply is made by the drop of the power stage’s anode current across \( D_12 \). Regulation of the high–voltage supplies is not used because the feedback across the power stage stabilizes the output voltage in spite of power supply variations.

A floating power supply consisting of \( TR_1 \) and associated components is used for the screen grid and heater of \( V_1 \). The ac heater voltage of 40V and the screen grid voltage of 100V can be generated from a single 40V transformer and a voltage-doubler rectifier (D1, D2, C3, and C4). The dual choke \( L_1 \) prevents the capacity of \( TR_1 \)’s secondary with respect to ground from loading the output by providinbg a high RF impedance. \( R_{24} \) critically damps the series resonance of \( L_1 \) and the transformer capacity. A similar (non-floating) power supply (\( TR_2 \) and associated components) produces heater and screen grid voltage for \( V_2 \). The transistor circuits are also operated from this supply.

\( T_1 \) and \( T_2 \) act as shunt voltage regulator to provide an adjustable 42V bias voltage at the cathode of \( V_2 \). It uses -5V from the power supply as a reference.

With 400Vpp sinusoidal output voltage at 3 MHz and \( C_l = 140 \text{pF} \) (plus 30pF wiring capacities), the power stage draws about 200mA average current. Under such
TABLE IV: Rise and fall time \( t_r \), \( t_f \), and 3dB-bandwidth \( B \) for small (\( \leq 200 \text{Vpp} \)) and large signals (400Vpp). Phase shift is about 90° at the -3dB frequency. With a 250 pF load, the amplifier will be about 25% slower than with 140pF.

| \( V_{pp} \) | \( U_2 \) | \( C_t \) | \( t_r \) | \( t_f \) | \( d \) | \( B \) |
|---|---|---|---|---|---|---|
| 200 | 20 | 40 | 55 | 40 |
| 400 | 20 | 57 | 60 | 50 | 7.5 |
| 200 | 140 | 45 | 90 | 55 |
| 400 | 140 | 60 | 70 | 70 | 5.7 |

TABLE V: Technical data of the 550 Vpp amplifier.

| Parameter | Value |
|---|---|
| Output voltage range | 50-600V |
| Input voltage range | \( \pm 2.5 \text{V} \) |
| Gain | 100 |
| Input impedance | 1kΩ |
| Load | up to 250 pF |
| Noise (refr’d to input) | 10nV/\( \sqrt{\text{Hz}} \) |
| Hum | 0.5 Vpp |
| Harmonic distortion* | 1.3% |
| Max. peak output current | \( \pm 800 \text{mA} \) |

*Measured at 400Vpp with a 140pF load at 1MHz.

dissipated power is removed by a small fan.

The QQE06/40 vacuum tube [13] is chosen for its low output capacity for a tube of its rated power, and sufficient transconductance of about 20 mA/V at the operating point (set by R10), giving the amplifier a voltage gain of 16 dB. The maximum output voltage is about 140 Vpp with a supply of 180 V. This is reached with about 30 dBm, or 20 Vpp drive level, that can be taken from a commercial semiconductor amplifier.

R3 and R4 prevent parasitic oscillations that might occur at UHF frequencies. It is crucial that C2-4 are connected with minimum lead length from the cathode to the end of the coaxial cable bringing in the signal, rather than to the housing. Otherwise, the resonance of this lead length would decrease the gain strongly above 100 MHz.

The circuit utilizes a LC compensation network to increase the bandwidth from about 1/(2\( \pi R_a(C_a+C_{ext}) \)) = 26MHz (where \( C_a = 6.4 \text{pF} \) is the output capacity of the QQE06/40, \( C_{ext} = 12 \text{pF} \) is the sum of the EOM and wiring capacities, and \( R_a = 330 \Omega \) is the anode load resistor) to 120MHz [12]. The inductance of the paralleled wire-wound resistors R5-7 as well as L1 compensate for the gain loss at high frequencies due to the circuit and EOM capacities (both are about 0.9\( \mu \text{H} \)). L1 is made of 10 windings of 0.3mm thick copper wire around the 1W metal film resistor R8, which removes a sharp spike in the frequency response that would occur with a lossless inductance here. The LC compensation as shown is optimized for a high bandwidth (140 MHz, see Fig. 6), but leads to some overshoot in the square-wave response and a large phase-shift of 90° at 40 MHz. For some applications, like driving an actuator in phase locked loops, a low phase shift is more important. The circuit may also be optimized for low phase shift by setting L1 to zero and increasing the inductance of the anode load resistors.

B. 140V, 120 MHz driver

Fig. 5 shows an amplifier capable of driving a low-capacity (10 pF) electro-optic modulator at 140 Vpp with 120 MHz bandwidth. The unit is built in a small housing that can be directly attached to the EOM, eliminating the cable capacity that would reduce the bandwidth. The
FIG. 6: The frequency response of the EOM driver is 3 dB down at 140 MHz.

TABLE VI: Technical data of the 120MHz, 140Vpp amplifier.

| Parameter            | Value                     |
|----------------------|---------------------------|
| Output voltage       | 140Vpp                    |
| Input voltage        | 20Vpp                     |
| Voltage gain         | 7                         |
| Input impedance      | 50Ω                       |
| Load                 | 10pF                      |
| Short circuit duration| Infinite                 |
| Bandwidth            | 0.01-120MHz$^a$           |
| Phase shift (40MHz)  | 90°                       |

$^a$may be DC coupled by small modifications

The technical data of this amplifier is summarized in table VI. The load used for this measurement was 10 pF connected via an SMA connection that adds about 2 pF.

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[3] Rise time: time between the 10% and 90% points of the rising edge of a square wave; fall time: time between the 10% and 90% points of the falling edge of a square wave; delay: time required for the output voltage of an amplifier to reach 50% of its final value relative to the instant at which a square pulse of negligible rise time is applied at the input. (3dB-) Bandwidth: frequency where the output amplitude of an amplifier driven by sinewaves of constant amplitude is reduced to $1/\sqrt{2}$, or -3dB, of its maximum.
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[10] BFQ252 and BFQ262 data sheets (Philips Electronics, Eindhoven, The Netherlands, 1997). Similar transistors are made, e.g., by the Toshiba and On semiconductor companies.
[11] The available peak current is given by the tube type used and the screen grid voltage. In this circuit, above 800mA are reached with PL509 tubes as specified in the original data sheets [12, 13], but some tubes of present day construction provide lower peak current and are thus not suitable.
[12] PL509 and PL519 data sheets, in: Empfängerröhren (Valvo GmbH, Hamburg, Germany, 1971).
[13] See, e.g., [http://frank.pocnet.net/index.html](http://frank.pocnet.net/index.html)