EER PA for 80 – 15 m

400 Watts out of only one transistor

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Today's commonly used class AB/B power amplifiers do have an efficiency of 50...60%. Larger power amplifiers are only possible by combining a number of modules. The efficiency of such combinations is degraded to around 30% and cooling of such voluminous constructions becomes a problem.

Already at the beginning of the 50s of the last century R. Kahn (1) published and patented a system of envelope elimination and restoration (EER) in the USA. As shown in **Bild 1** the SSB signal is splitted into the envelope part by the envelope detector (HK Detektor) and into the phase portion by the limiter (Begrenzer). The phase portion then is limited in such a way that practically any modulation of the amplitude is eliminated. By this it is possible to reach a high degree of efficiency with power amplifiers. The SSB linearity of the PA tube or the PA transistor does not play any role anymore. The task of the modulator is to re-modulate the envelope curve delivered by the envelope detector back onto the power amplifier.

Pulse-Density-Modulator as the way out

If a class B or AB amplifier is taken nothing will be gained regarding efficiency – the problem of the high power dissipation only is shifted onto the modulator. The pulse-density-modulator (PDM) is the way out. It works similar to a switched power supply. Although the EER principle is known since decades mainly the AM envelope modulation has been implemented in high power broadcasting amplifiers. There they are achieving big power cost savings up to six digits since many years. The EER principle for SSB is more complicated and this might be the reason why it cannot be found to a large scale in commercial as well as ham equipment. Unfortunately there is also not much literature available. Especially complete systems for ham operation are missing. Therefore development of the single modules as well as of the whole system was extremely time consuming.
Within the past four years there have been various descriptions of so-called class E amplifiers in the american ham literature [2, 3, 4]. For the first time it was introduced in 1975 by Nathan and Alan Sokal [5].
Described is a switched amplifier with an efficiency of 90% and more. The power transistor resp. the power tube is not a current driven source like the usual class A-, B- or C amplifiers. It works as a switch. In **Bild 2** the switch S1 refers to a power stage transistor. It is opened and closed for each half of a period by the RF control stage. The reactance of the RF choke prevents a DC short and at the same time the choke stores the current which was flowing through the switch and after opening flows through the coupling network C2, C1 and L1 into the load (there are similarities with a level shifter of a switched power supply). **Bild 3** shows the voltage at the transistor gate (Ch 1) and at drain (Ch 2) over time. The reason for the high efficiency is that the current-times-voltage multiply of the transistor is small during the RF period. Assuming the correct dimensioning of the output network the transistor is switching at the current/voltage minimum and

therefore has a very low power dissipation. Furthermore it can be seen that this is a narrow band circuit. This is not a disadvantage in ham radio because even with broadband solid state amplifiers there is the necessity of a low pass filter for every band. In its basic construction the class E amplifier is only suitable for CW and FM/FSK.
Besides the high efficiency there are other advantages as well:

- because the transistor acts as a switch the non-linearities of its characteristic do not apply to the output signal
- in contrary to current driven power stages the match of a class E amplifier always is optimal. The current results from the momentary envelope voltage of the PD modulator divided by the transformed antenna impedance (U_{hk} / R_{ant}). So, the straight line characteristic is achieved automatically.

Envelope conversion and control of the pulse-density modulator

In order to achieve the highest possible degree of efficiency the principle of pulse-density modulation was used for the modulator. Therefore the AF envelope (NF Hüllkurve) first of all has to be converted to a squarewave signal by a differential comparator (Vergleicher) (**Bild 4**). Then a power FET is switched after the signal has passed through an isolation coupler and driver. The low pass filter at the FET output converts the squarewave impulses back into a voltage which is proportional to the envelope curve and which is fed into the RF amplifier.
When dimensioning the modulator some conditions have to be considered:

1. the intermodulation products of the PDM
2. the bandwidth of the envelope channel
3. the group delay between envelope channel and phase channel

In order to achieve a reduction of unwanted intermodulation of -40dBc the sample frequency f_S according to [6] must have a ratio of >6.5 compared to the SSB bandwidth B_{hf}.

The bandwidth of an SSB envelope curve generally is indefinite. Therefore the finite bandwidth of the envelope modulator (HK-Modulator) distorts the envelope which is represented as IMD in the output signal. The bandwidth of the envelope channel is limited insofar that a practical ratio between envelope channel and sampling frequency must be obtained. Otherwise it will be very difficult to sufficiently surpress the sampling frequency in the output signal.
According to [6] for -40dB IMD the maximum allowance of group delays is up to 13.3µs. The low pass filter in the envelope detector (HK-Detektor) has a delay of 12µs. The envelope control including the low pass filter within the PDM together add another 25µs . Counter measures to compensate for the group delays therefore are essential. Also Dr. Karl Meinzer, DJ4ZC, many years ago has shown graphically in [7] the distortions of EER systems in relation to bandwidths in the phase and envelope channels. The purity of the output signal therefore is additionally determined by the envelope detector and the linearity of the modulator. Theoretically only the demodulation distortions of the envelope detector should exist if the group delay compensation between phase channel and envelope curve channel is sufficient.

The details of the envelope curve PA

The block diagram (**Bild 5**) shows all relevant stages and construction modules of the system. Within my homebrew transceiver SSB is generated with a 200-kHz filter. In order

to achieve the lowest possible demodulation distortion the SSB signal level is amplified to 14 V_{ss} through an OP LF357 (**Bild 6**). The following AF lowpass filter has a limiting frequency of 30 kHz and a constant group delay of 12µs up to 10 kHz. The amplified SSB signal then is delayed by 20µs within a 200-kHz-allpass filter, amplified by 3dB with an additional LF357 and then limited to 0.8 Vss by the two Schottky diodes BAT41. The already existing construction modules transform the signal up to the final frequency. This path has a group delay of 17µs (**Bild 7**). Such, together with the 200 kHz allpass filter the group delay of the envelope channnel of 37µs will be compensated for in the phase channel. The output signal of the broadband amplifier still carries portions of the envelope curve despite the first envelope curve limiter. These must be eliminated with help of additional limiters. This necessity is obvious when looking at an SSB two tone signal. At the constrictions it lowers to zero. If there is unsufficient limitation then the drive for the final stage is missing at these nodes.
The envelope control stage as well as the envelope comparator stage can be seen in (**Bild 8**). The constant current source Tr2 is charging capacitor C8. After reaching a certain value the adjustable precision Schmitt-Trigger – comprising of OP4 – is discharing C8 through resistor R13 und flips back into its initial state after reaching its lower threshold. The generated 100 kHz triangle drives the negative input of the differential comparator OP2. The 100 kHz sampling frequency can easily be removed through a four stage lowpass filter at the output of the PDM. Through an RF filter OP1 receives at its negative input the AF envelope (negative polarity). This node also receives the other envelope (positive polarity) which comes from the RF power amplifier. This feedback loop prevents the PDM from being driven into saturation. The controlled envelope signal is then converted into 100 kHz square wave impulses within the comparator OP2. Their duration matches the momentary value of the envelope curve. Both of the NAND gates care for the necessary polarity in order to drive the optical wave guide TOTX173.

Driving as in switching power supplies

Driving the pulse-density-modulator must be done with ground floating – as in switching power supplies. Contrary to switching power supplies the amplitude variation as well as the slew rate of our SSB envelope are remarkably larger. In order to achieve an IMD of -40dB the envelope amplitude must be resolved at least by a ratio of 1/100. This means that also the pulse duration has to vary within the same ratio. 1% of our time frame of 10µs (at 100 kHz sampling rate) now only means 100ns. Additionally one needs rise times (t_{rise}) of at least 30ns in order to be able to realize a sufficient 100ns pulse. The required bandwidth now is $0.35/t_{rise}$ – appproximately 12 Mhz. The fastest optocouplers available today cannot be used here. The optical wave guides TOTX173/TORX173 from Toshiba are a loophole here but with a data rate of 6 Mb/s they are far from a perfect solution. The Analog Devices AD261 surely is the right choice for later. Also the rise times and fall times of the driver as well as the MOSFET switch within the PDM are problematic in this regard (**Bild 9**). A way out would be to lower the sampling frequency to 50 kHz. But then we come close to the limiting frequency of the integration lowpass filter within the PDM. The surpression of the sampling frequency then would not be that easy anymore.
An isolated supply voltage is necessary to run the driver. To be able to use commonly available N-channel-MOSFETS the negative voltage is switched. The integration lowpass consists of two RM-14 pot cores and is dimensioned for a frequency of 20 kHz and an impedance of 20 Ω.
For the class E amplifier an RF MOSFET type ARF447 from Advanced Power Technology is used. The transistor has a power dissipation of 230 Watts (at 25 °C) in a TO-247 package.

The circuit can be seen in (**Bild 10**). For the dimensioning of the matching network I first used the values in [5]. But they are only successful up to 40 m. From 20 m onward there was no square wave pulse response possible at the drain. The reason is the explosively rising output capacity (C_{OSS}) if the drain-source-voltage is lowered to 0V. The ARF447 e.g. shows a rise from 100pF (at 100 V) up to over 1000 pF (at 0 V). This effect is especially pronounced with a switched amplifier because contrary to a current driven linear amplifier there is no saturation voltage available anymore. Simulations of the matching at manually varied drain voltages showed matching drops up to several Mhz.
The ARF447 has a breakdown voltage of 900 V and is designed for up to 300 V supply voltage. This gives comfortable working impedances of approximately 140 Ω which unfortunately cannot be used due to the high C_{OSS}. Experiments with a series tuned circuit between the transformation coil and the first coil of the 50 Ω lowpass filter showed success. The L of the series tuned circuit is not critical, the series tuned C must be adjusted to a square wave pulse response. However, the work impedance for the ARF447 had to be reduced to 20 Ω which after the SOA diagram can be tolerated for SSB and 100 V supply voltage. The matching networks are assembled on a double-sided PCB of the size 230mm X 145 mm and are toggled with high current relays. The parts have been soldered on pads which were milled out of the PCB. For the capacitors only mica or strong ceramic types should be used with a minimum of 500 V DC. Most of the coils are air coils wound with 1mm or 1.5 mm copper wire or silver-copper wire. The series Ls of the 80m and 40 m lowpass filters have been wound on old Siemens ceramic bodies with spindle core. The relays are located at the bottom side of the RF PCB.

9:1 ratio RF broadband transformer needed

The RF transistor is driven through a 9:1 ratio RF broadband transformer. Due to the high gate capacitance of 2nF a transformation to 2.8 Ω of load impedance has to be done. In order to be able to drive the gate to at least 6 V_{peak} a driving power of 13 W is necessary. The stray inductance of the secondary coil (two brass pipes through two 80K1 binocular cores) is compensated with C2 (mica). The RF broadband transformer when compensated has a straight line frequency response up to 30 Mhz with a return loss of -20 dB. Caused by the C load of the RF FET it acts as a lowpass from 22 Mhz onward so that 12 m and 10 m cannot be reached anymore. It is up to further investigatons if a switchable band matching network with bundled elements could get us on.
The schematic of the phase driver can be seen in (**Bild 11**). In order to be able to also feed higher power into Bu2 we use a 50 Ω / 25 W chip resistor for termination. Both of the Schottky diodes BAT41 are limiting the envelope curve fractions being present to 0.8 Vss. The main limiting is done within the differential comparator NE521. This one is reacting on an input voltage difference of 5 mV with a TTL jump at its output. After this the phase channel is free of any envelope curve fractions. The three stage AB broadband amplifier is frequency compensated and delivers a maximum of 40 Watts up to 40 Mhz. In order to simplify the transformation ratios a 25 Ω output transformer was chosen. Due to the high total limiting within the phase channel it had been reasonable to lower the subcarrier within the transmitter only to -40 dBc. With a fully surpressed carrier left and right noise peaks are visible in the spectrum while the modulation is paused.

400 W out

The envelope PA delivers an output power of 400 W (PEP) on all bands. The efficiency of the class E amplifier is 90%. I do not know of any solid state PA which delivers 400 W with just one transistor. (**Bild 12**) shows the spectrum on 80 m with a two-tone test. The right marker indicates the two-tone levels in the upper sideband and the left marker indicates the IMD3 level in the lower sideband. The IMD3 span is 33.3 dB which gives 39.3 dB relative to the PEP power. The spectrum of the upper sideband even looks better. I think the results are respectable. The values at 40 m are 3 dB less. Unfortunately the linearity goes down with increasing frequency. On 15 m the IMD3 span only is -16.2 dB (**Bild 13**). The reason is a phase modulation caused by the varying output capacitance C_{OSS} of the ARF447 which becomes more and more effective with increasing frequency. Surprisingly enough, the spectrum within the unwanted sideband sounds rather monotone. The normally present agressive peaks are missing here. The vertical lines in the spectrum (**Bild 14**) are marking the bandwidth of the SSB filter and it can be seen that the energy within the lower sideband decreases very quickly to below -40 dB. There is hope that the new Cool-MOS-Power transistors from Infineon with their reduced output and feedback capacitances as well as the new SMPS-MOS-FETs from IRF will deliver better results. The measured harmonic surpressions are at -46 dBc for the first harmonic at 20 m and 17 m. All other bands are performing better, also with higher order harmonics. Experiments have been done with two switching power MOSFETs of types IRF710 and IRF820 respectively in parallel and they were showing even better IMD3 results on 80 m and 40 m (almost -40 dB based on single-tone). But from 20 m onward they get hot only by the RF driving power. Obviously the chip structure is not suitable for RF use. With two of those MOSFETs a maximum of 200 W PEP can be reached with reduced supply voltage and with the same output network (tested up to 20 m). Because both transistors will never have the same drain-source resistance (R_{DSon}) the better one of the two always 'dies' first. The switching frequency shows a surpression of -60 dB within ±100 kHz of the output signal but can be improved up to -80 dB with an additional 100 kHz pole (9 nF across L2) within the PDM. The envelope shows an astonishing robustness against mismatch. Even with an SWR of 3 it remains symmetrical.

Experiences with tubes

Tests have been made with a horizontal deflection amplifier tube EL519 on 80 m and 600 V supply voltage. The weakness of this tube is its current consumption which is heavily reduced below its knee voltage. With this the PDM will operate in the so-called gap mode. If the quiescent voltage is adjusted to 50 V there is not enough resolution of the envelope anymore. The ratio of voltage supply to knee voltage is 12 which is around 22 dB. This also is the value which can be achieved as IMD3 span. Compared to the classic AB amplifier nothing is gained except for some improvement of the efficiency. It still needs to be investigated if there are performance improvements possible by certain load compensation tricks or with a class C amplifier stage. Also there is hope that on the higher bands better IMD values can be achieved due to the far lower tube capacitances.

Summary and prospects

The report is meant to deal with my experiences gained on EER over the last four winter seasons. The EER principle gives some advantages over normal power amplifiers like

- significantly improved efficiency
- less space
- less cooling problems

- better linearity
- lower power dependency

I can only hope that also other OM get interested in this topic and will achieve some improvements with new ideas. This is especially valid for those who know how to use modern DSP techniques. With this the hardware complexity could possibly be improved as well as the technical performance. Almost daily I can be met in the ham bands operating my envelope PA and I am looking forward to any technical discussion about this topic.