The "Stuttgarter-Geigerle"

Version 1 Bernd Laquai 31.5.2012

For this variant of pin diode counter I chose the name "Stuttgarter Geigerle" according to the habit in my hometown Stuttgart to attach the suffix "le" to belittle all what is valuated, small and cute. The "Stuttgarter Geigerle" is a spatially very small construction of a Geiger counter on the basis of an transimpedance amplifier (TIA) followed by a comparator and a red/green toggling LED indicator. The circuitry and the mechanical construction is fairly small in complexity compared to other solutions and is therefore very simple and low-cost for a do-it-yourself instrument without losing value and functionality. The measuring amplifier follows in general an application note for a photo detector from the former semiconductor company Burr-Brown (in the meantime part of Texas Instruments). The original paper can be found here:

http://www.ti.com/lit/an/sboa035/sboa035.pdf

The circuitry of the "Stuttgarter Geigerle" particularly follows fig. 4a of this document.

Basics of the circuitry

Transimpedance amplifiers

Today's operational amplifiers are circuits that amplify signals in terms of voltage. However, it is a basic property of a PN- oder PIN semiconductor junction, that when operated in photovoltaic or photoconductive mode, it behaves like a current source with an almost infinite impedance. This means when a radiation quantum strikes that junction, whether this is visible or radioactive radiation, in will generate a current impulse rather than a voltage impulse. Provided the radiation originates from a natural source, such a current pulse has the size of a few nanoamperes per sensed disintegration.

To post-process this signal it is required to transform it into a voltage signal. One approach is to force this current flowing through a very large resistor to generate a voltage drop as large as possible for further amplification. One problem with this approach is that the voltage depends on the size of the current impulse in a logarithmic way as a consequence of the exponential characteristic of the PN junction.

The other option is, to add circuitry to an operational amplifier in a way, that it amplifies the current directly and outputs it as a voltage. When the amplification is defined as the ratio between output voltage and input current, then this ratio has the property of a resistor. Assuming a sinusoidal signal with amplitude and phase the dimension of this ratio is called impedance. Therefore, a 1nA input current generating a 1V output voltage, causes a "transimpedance amplification" of 1 Gohm. To amplify the input current linearly, the voltage at the diode junction must not change. With respect to the fact that the reverse current (dark current) which is radiation independent can be cancelled it would be preferable to have this voltage being constantly zero. However, with respect to the stability of the circuit and the sensitivity for fast pulses it is advantageous to reversely bias the diode while accepting a reverse current. When keeping the reverse bias at a constantly high voltage it

helps minimizing the voltage dependent capacitance of the junction and thus increases stability and the sensitivity for fast pulses.

In case of a transimpedance amplifier the detector diode is connected directly at the negative input. Only a feedback resistor Rf in the feedback loop causes a decreasing output voltage until the current through the diode is completely cancelled by the current through Rf in the regulation loop. This regulation loop is designed to force the difference between the positive and the negative input terminal of the OP to zero. When the feedback resistor is large, the voltage drop to cause the cancellation of the small diode current is also large respectively. A prerequisite for the functionality of this principle is an almost infinite input impedance of the operational amplifier.

Ideal model of the transimpedance amplifier

In its most ideal form an operational amplifier can be modeled as a controlled voltage source amplifying its input voltage with a very large gain A (e.g. A=1E6). When this ideal source is fed back to the negative input the output will counteract the input voltage difference until this vanishes. Therefore an input voltage difference of zero can be assumed in the equilibrium state. This will ease all calculations significantly.

This means that when the positive input is at zero voltage then the negative input will also be close to zero. From this it can be concluded that the output voltage completely drops along the feedback resistor. Assuming additionally that no current flows into the amplifier input due to very high impedance (and low input leakage) then the diode current and the current through the feedback resistor must be equal. Or in other words the output voltage is equal to the input current times the value of the feedback resistor. Hence, the feedback resistor value directly determines the transimpedance amplification value. Only the factor -1 has to be taken into account additionally, since the feedback is working on the negative input.

If one is interested in the calculation in more detail, the approach is as follows:

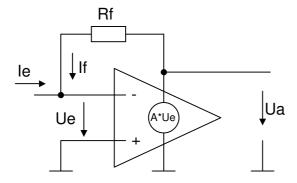


Fig. 1: Schematic functionality of a transimpedance amplifier

According to Kirchhoff's loop rule it is:

Ua = If*Rf + Ue

Under the condition that the input leakage current into the OP is much smaller than Ie, Kirchhoff's point rule requires that:

If=-le

Assumed the open loop gain of the OP is A it follows:

Ue = Ua/A

Both above results fed into the loop rule, this yields:

Ua = -le*Rf + Ua/A

Solving this equation for Ua the following is obtained:

Ua = -Ie*Rf / (1 - 1/A)

Now it can be quickly recognized that already for A > 1000 it is valid to assume:

Ua = -le * Rf

For real OPs we can assume that A > 100000. In other words, we can approximate the transimpedance amplification quite accurately with T = -Rf.

For large A we can also assume that Ue is almost zero and the input node is always constantly close to zero volts, independently of the magnitude of Ua. Therefore when connecting the diode detector to the negative input we can expect a linear dependence of the output voltage to the input detector current due to the constant current. Since the diode current is also proportional to the energy of the sensed radiation quantum the output voltage is also proportional to the radiation energy striking the active area of the detector.

This linear relationship is not present when the diode current is first converted into a voltage at a shunt resistor and then being amplified using a conventional voltage amplifier. In such a case the voltage at the shunt is logarithmically dependent on the detector current and therefor on the radiation energy.

The implementation of the circuit

Input leakage and amplifier gain

A prerequiste for an OP to be used as transimpedance amplifier is an input leakage and impedance that is one order of magnitude smaller than the current to be amplified. So when a current impulse sourced by the detector diode is in the range of a nanoampere (and the dark current is in the same range) the input leakage current should be less than some 10th of picoamperes. An OP that uses a bipolar transistor input stage will not achieve such low leakage (bias) current values. Either junction FETs or MOSFETS need to be used in the input stage. When great care is taken and e.g. guarding techniques are used transimpedance amplification of 100GOhms are possible with today's leading edge OPs.

Bandwidth and Noise Power Density

The bandwidth given for OPs normally is the bandwidth obtained for unity gain (it does not mean a OP is stable for gain 1). This bandwidth is also called unity gain bandwidth or gain bandwidth product. To calculate the real bandwidth for a voltage amplifier for a given gain the unity gain bandwidth in the datasheet has to be divided by the target gain of the amplifier. This is not required for the transimpedance amplifier circuit in its ideal form. Ideally one can observe the true unity gain bandwidth. For a transimpedance amplifier the real bandwidth is only limited by the parasitic capacitances, predominantly by the capacitance of the detector diode at the input.

For this reason it is recommendable not to operate the detector in photovoltaic mode (without reverse bias) but in photo conductive mode with a reverse bias. A reverse bias will empty the PIN junction from all charge carriers which in turn reduces the capacitance dramatically. The disadvantage is that the reverse (dark) current no longer will vanish. However, for a good photo detector diode it will be still less that the current generated from the sensed radiation. Taking the reverse bias into account the input circuitry looks like this:

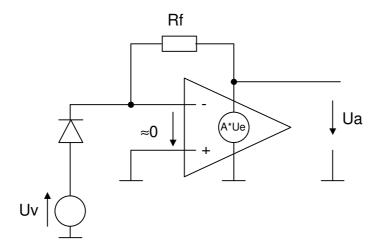


Fig. 2: Transimpedance amplifier as photo detector amplifier in photoconductive mode

The noise in a transimpedance amplifier is mainly caused by the thermal noise from the large feedback resistor Rf (E=4*k*T*Rf, k Boltzmann constant) as well as by the current noise of the amplifier itself (given in pA/ sqrt Hz). Unfortunately the current noise is often not provided in the datasheet by the semiconductor manufacturers of an OP. The additional term of the current noise that adds in the RMS sense to the thermal noise is given by In*Rf (In = current noise referred to the negative input). It quickly becomes evident that a view pA of current noise can cause several mV at the output. If furthermore this value is multiplied with the required bandwidth of at least 20kHz it can be concluded that the output noise easily reaches several 100th of microvolts at the output when the wrong OP is selected. However, OPs with junction FETs or MOSFETs are available where the current noise is in the range of femtoamperes rather than picoamperes and only the thermal noise becomes visible. On the other hand it becomes also visible that it does not make sense to provide an unlimited noise bandwidth, because in such a case more noise energy is amplified compared to the signal energy.

Single voltage power supply

Often OP circuits are designed in a way that the reference voltage (ground) is at OV and two symmetrical power supplies +Vs and –Vs are used. Such architectures generate an output voltage that can either be above or below zero volts. However, the circuit overhead is higher than for a single supply voltage. Therefore it would be advantageous to manage the power supply from just one voltage source. This can be achieved by moving the reference voltage to a voltage produced from a voltage divider between ground and a single positive supply. To reduce the high frequency noise the voltage divider should be sufficiently blocked with a capacitance. In this arrangement the reverse bias for the detector diode is also achieved easily.

It would be natural to position the reference to half of the single power supply voltage. However, since the capacitance of the detector diode is the smaller the higher the reverse bias is, it is preferable to move the reference voltage of the first stage to 3/4th of the operating voltage. Since the DC portion of the signal needs not to be amplified (the dark current) the following stage can be coupled with a capacitance and the reference for this second stage can be moved back to the half of the operating voltage.

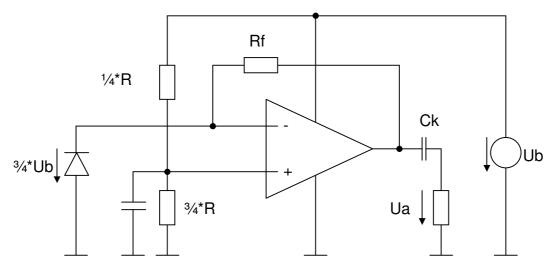


Fig.3: Supply of the transimpedance amplifier with only one supply voltage

Stability

Since the transimpedance amplifier is very sensitive to small currents that are amplified with very high gain, there is a risk for crosstalk of the output signal to the input. As long as the crosstalk is not inversely amplified (phase shift of the signal by 180 degrees) it is not a big issue. However, if there is a frequency for which the signal is amplified with a 180 degree phase shift, oscillations will start rendering the circuit useless.

The internal junction capacitance Cd of the detector diode exactly is responsible for an increasing phase shift with the increasing frequency of spectral components contained in the signal. In the same way the limited bandwidth reduces the gain with increasing frequency. If a spectral component for which the phase shift is 180 degrees is still amplified with a gain greater than 1, the circuit becomes instable and starts to oscillate.

In general this tendency for oscillations caused by the diode capacitance can be avoided when the ohmic feedback resistor Rf is bypassed with a small capacitance reducing the amplifier bandwidth. However such a measure may also reduce the signal bandwidth adversely attenuating the size of the radiation impulses. The capacitance for the compensation therefore needs to be very small (often less than 1pF). As a consequence the PCB layout and size of the feedback resistor may have a significant impact.

Since it is very difficult to get precise capacitances below 1pF and generating an appropriate layout a tricky circuitry is required to achieve the right compensation goal. Of course it would be possible to put some larger capacitances in series to form a smaller one. On the other hand it would be advantageous being able to precisely adapt the capacitance to the specific requirements of an implementation influenced by the layout and part variations with an adjustable value. To address this issue Burr-Bown provided an excellent proposal in the above mentioned application note. It uses two large capacitances C in series and a trimmer capacitance Ct in the junction node between connected to ground with the other terminal. This circuit acts like a capacitive voltage divider only passing along a portion of the high frequency feedback signal to the input. Therefore the effective capacitance is the smaller the larger the capacitance of the trimmer capacitance is adjusted (in comparison to the two series capacitors). This also allows to easily adapt to different detector diodes and to manufacturing variations of a selected OP.

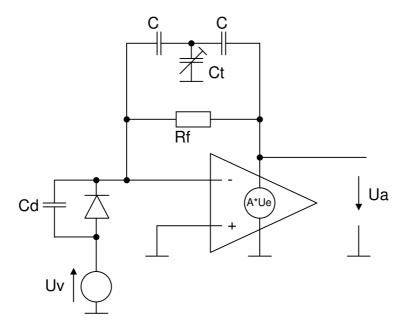


Fig. 4: Adjustable compensation of the detector junction capacitance

The second amplifier stage

When using regular PCB and assembly technologies it is recommendable to keep the feedback resistor smaller than 10Mohm. If you go for more parasitic leakage may quickly destroy this trial. If we assume a 1nA peak of the detector current impulse a 10Mohm transimpedance will produce an output voltage of 10mV. To do a solid comparison if a

radiation quantum was sensed and not just a noise excursion a signal peak of about 1V at the output of the whole amplifier should be reached. Therefore a second stage with an amplification of about 100 is still required. Because many OPs are available as a "dual-version" in one package this is possible with almost no additional effort. Since the output of the first stage is already fairly low in impedance a conventional inverting voltage amplifier stage can be used in conjunction with a reference voltage in the mid of the supply voltage.

In practical studies however, it turned out that the second stage needs also be compensated by a small capacitance in parallel to the feedback resistor to limit the bandwidth and with it also the noise and thus to increase stability.

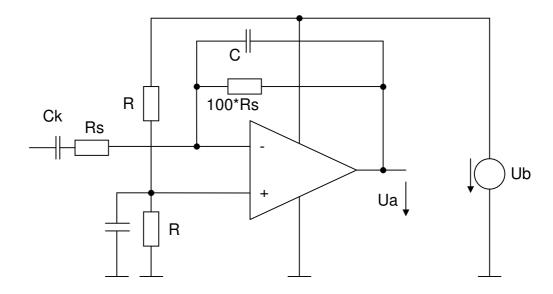


Fig. 5: The second amplifier stage

Selection of Parts

As a PIN detector diode the BPW34 from Vishay (and other manufacturers) are such cheap and good that I focused very quickly to this product as a detector. It has an active area of 7.5mm² what is of course very small compared to a Geiger-Mueller tube. Respectively, the counts per second are smaller for a source with a large radiation area however, a much better spatial resolution is obtained to detect small hot spots, impurities or pollutions in small samples. The capacitance is 70pF at 0V and only 25pF at 3V. This means a reverse bias really creates a big improvement. To achieve acceptable counts for natural samples that were available to me, (e.g. granite stones from the black forest area) I put three BPW34 diodes in parallel. I believe it will be possible to put up to 5 PIN diodes in parallel, however loosing somewhat of the spatial resolution and reaching saturation more quickly for highly radioactive samples.

The central part is the OP that needs some care for selection. In most of the cases an experiment with a prototype is required to check the suitability of an OP. The ideal model based on controlled sources is often not sufficient for this purpose since there are too many parasitic sources that also determine the behavior of the amplifier. The parasitic elements however are hardly to predict.

I obtained very good results with the J-FET OP from Analog Devices AD8666. It has an input leakage current of less than 1pA, a bandwidth of 4MHz and is marketed as a "low-noise" amplifier with a voltage noise density of 8nV/sqrt Hz referred to the input. The current noise density unfortunately is not specified but I assume it's negligible to all other noise sources. The OP part is available for less than 3Euro (status May 2011). Using three BPW34 in parallel as detector I was able to achieve stability with two 10pF capacitors in series and a 200pF trimmer at minimal position. The OP works with a single supply voltage of up to 16V. This allows reaching a significant reverse bias for the detector diodes.

I was able to determine the frequency characteristic of the circuit with a weakly forward biased red LED diode modulated with a broadband white noise and the amplifier positioned in a metal enclosure protected from any light and radiation. Using a feedback resistor of 10Mohm, three BPW34 PIN diodes as detector and the 200pF trimmer in minimal position the resulting 3dB bandwidth was about 26kHz.

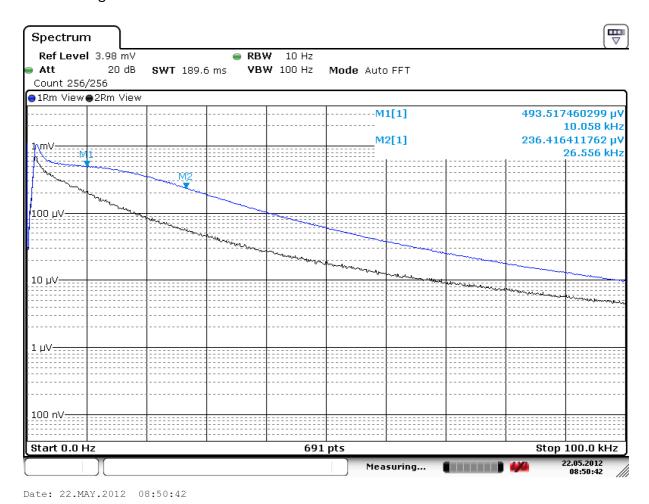


Fig. 6: Frequency characteristic of the amplifier measured with white noise visible light. Blue curve: trimmer OpF, black curve: trimmer 200pF

I coupled the second stage to the first one using a 100nF capacitor to suppress the DC portion of the signal which is of little interest. In the second stage the gain and feedback resistors were selected to be Rs=1k und Rf=100k resulting in a gain of 100. With this design an old wrist watch with radium painted pointers yields pulses of up to 2V amplitude at the

output of the second stage. The noise is below 20mV rms and the current consumption during operation is below 5mA when using the AD8666.

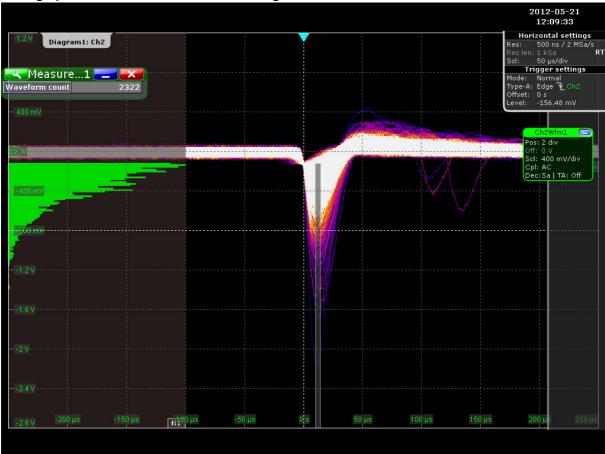


Fig.7: Amplified pulses for the radiation generated by radium painted pointers of an old wrist watch. The histogram (green) shows the distribution of the pulse heights (aluminum foil as detector cover).

The AD4001 from Analog Devices (very similar to the AD8666) also works well but I needed two times 20pF plus a trimmer value of about 50pF for compensation to get it stable. An experiment with the LTC6241HV CMOS OP from Linear Technology yielded also very good results comparable to those of the AD8066. This CMOS device is specified with even lower input leakage and less noise and it is pin compatible to the AD8666 and available in the same SOIC package.

Comparator and pulse stretching

Pulses in the size of 1V in presence of noise with less than 20mV rms are no challenge for being digitized. At a typical impulse duration of > 10us a low cost OP like the LM358 of National (part of TI today) does the job. It also does not contribute much to the power consumption. To position the decision threshold precisely enough a trimming potentiometer with at least 10 turns is required. This potentiometer needs to be embedded between 2 fix valued resistors to enlarge the adjustment range. Since the second amplifier stage inverts the pulses of the first stage the comparator threshold needs to be placed 50mV below the reference voltage, which is 50mV below half of the supply voltage. This adjustment ensures

the correct triggering by radiation pulses and excludes noise excursions from triggering the comparator.

To just achieve the comparator functionality neither a negative nor a positive feedback is required for a standard OP. Without additional circuitry the OP would have a gain equal to the open loop gain which is quite high. The output voltage is limited in such a case only by the specified operational range, which is a bit smaller than the supply voltages. At the comparator output the pulses reach amplitudes close to the supply voltage and with almost the same impulse duration as the detector impulses. The duration however is too small to be recognized neither audibly nor visibly by a human. But when a positive feedback is introduced from the output to the positive input of the OP via a capacitor Cm, the pulses can be stretched until they can be heard or seen. The capacitor Cm works as a shortcut in the first instance of time. It therefore shifts the threshold against the reference voltage such that the comparator does not switch back, even though the input signal returned to above the threshold as adjusted with the trimming potentiometer. After a certain time the feedback capacitance gets discharged and the stationary threshold is effective again. Cm and R result in a time constant which should be in the millisecond range to get the audible pulses that sound authentic for a Geiger counter.

Since the LM358 is also a dual OP, the second amplifier can be used additionally as a buffer and to invert again the pulses. After that stage the signal can be output via a series R and a series C to limit the current and to suppress DC directly to an ear phone or to another post processing unit.

Additionally it is recommended to implement a power supply from a primary or rechargeable battery and to filter the supplies additionally for both the amplifier as well as for the comparator ICs with a RC or even a LC filter to prevent interferences from the comparator to the amplifier during switching and to prevent oscillations.

Optical LED display unit

In a noisy environment or when walking through a flea market hunting for interesting samples it is very convenient to just have a simple optical indicator for radiation. The noise of a Geiger counter may indeed create nervousness among other people which may be difficult to be calmed down again. Since individual short light pulses are hardly to recognize it may be better to have two LEDs toggling in a complementary way. A red/green dual LED in a single 5mm package preserves space and is perfectly suited for this purpose.

Since the OPs used for the measurement amplifier and the comparator both work with variable voltages of up to 15V supply voltage and all reference voltages scale accordingly it is desirable to also have a display unit that is almost independent of the supply voltage and there is no need for a power consuming linear voltage regulator. A toggle flip-flop (JK-Flip Flop) with exactly these properties is available from the 4000er CMOS logic series. It works with variable supply voltages of up to 15V without issues and has its switching threshold at 50% independently of Vdd. With such a logic gate the two LEDs can be connected to Q and Qn and J and K can be set such that the two LEDs toggle for each sensed radiation quantum. The HEF4027 logic gate (NXP) contains two J-K flip-flops. When one is used the other one should be disabled to minimize power consumption. When operating the amplifier,

comparator and display unit together, the total power consumption is about 15mA (twice as much as the amplifier).

The Overall Circuit

The overall circuit for the "Stuttgarter Geigerle" is obtained by cascading of measurement amplifier, comparator and LED display unit. The circuit can of course be also operated without the LED display unit. When using only an earphone the power consumption drops to about 6mA.

As a power supply either a 9V alkaline battery block can be used or a 9.6V NiMH rechargeable NiMH block (8 cells). An even smaller is a solution based on 3 LiPo cells yielding 11.1 V. The great advantage of the circuit is that all threshold scale with the used supply voltage and need not to be readjusted again.

It is also very easy to keep a rechargeable battery charged from a simple and small 12V solar panel. This helps to become independent from primary batteries what might be very important in an emergency case when no batteries can be purchased anymore. In such a case a rechargeable battery ensures operation for a long time for an instrument that suddenly becomes vitally important.

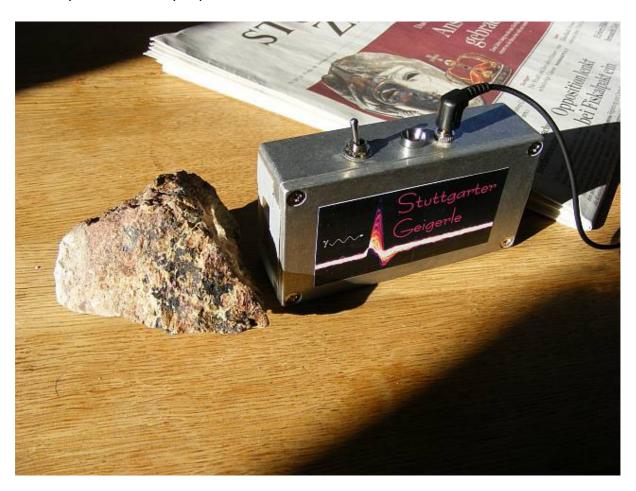


Fig. 8: "Stuttgarter Geigerle" with a granite sample containing small amounts of uranium salts from the Black forest area

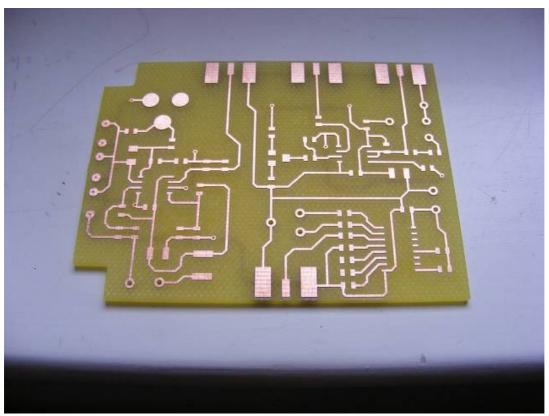


Fig. 9: Prototype PCB with additional test points



Fig. 10: A look inside the prototype



Fig. 11: Open detector aperture showing the 3 PIN diodes



Fig. 12: The detector aperture closed with adhesive aluminum foil

Fig. 13 Circuit diagram of the "Stuttgarter Geigerle"

