VINTAGE WORKBENCH

The Tektronix Type 130 LC Meter – Part 1
How it works

By Alan Hampel, B. Eng. (Electronics, Honours)

Unfortunately this sort of thing does happen. I was ripped off by a dodgy eBay seller – sold a bill of goods, you could say. But this story has a happy ending. I had a lot of fun converting a dirty, unusable relic into an as-new laboratory instrument with a rich history.

The T-130 LC meter from Tektronix was built from 1954 until 1975 and has five capacitance measuring ranges (3pF, 10pF, 30pF, 100pF and 300pF) with 1% FSD accuracy and a stable zero. Thanks to its 4.5-inch (~11.5cm) meter, it can easily resolve down to 0.05pF. It also has five inductance ranges from 3μH through to 300μH.

I bought it because I needed a capacitance meter that could accurately resolve sub-picofarad values for a project. I also collect and restore valve test gear, so the T-130 seemed like an ideal candidate. As such, one for sale on eBay caught my eye. The price was very reasonable, and it looked clean and original in the photos, so I bought it.

The seller claimed he had run it for a couple of days with a 25pF capacitor, and got a correct stable reading.

When it arrived, the package was not damaged, but turning it over produced clunking sounds. That's a bad sign!

As it turned out, the instrument was generously coated inside and out with cigarette smoke residue, and was inoperative due to many faults.

The origin of the T-130

During Tektronix's early days (see the side panel for a brief history), they needed an instrument to measure small capacitances, eg, stray wiring capacitance and valve capacitances, as well as small inductances. The production lines needed a stable instrument, usable by semi-technical operators. The lab needed accuracy and sub-picofarad sensitivity.

After joining Tek in 1951, young engineer Cliff Moulton designed the T-130 to meet just these needs.

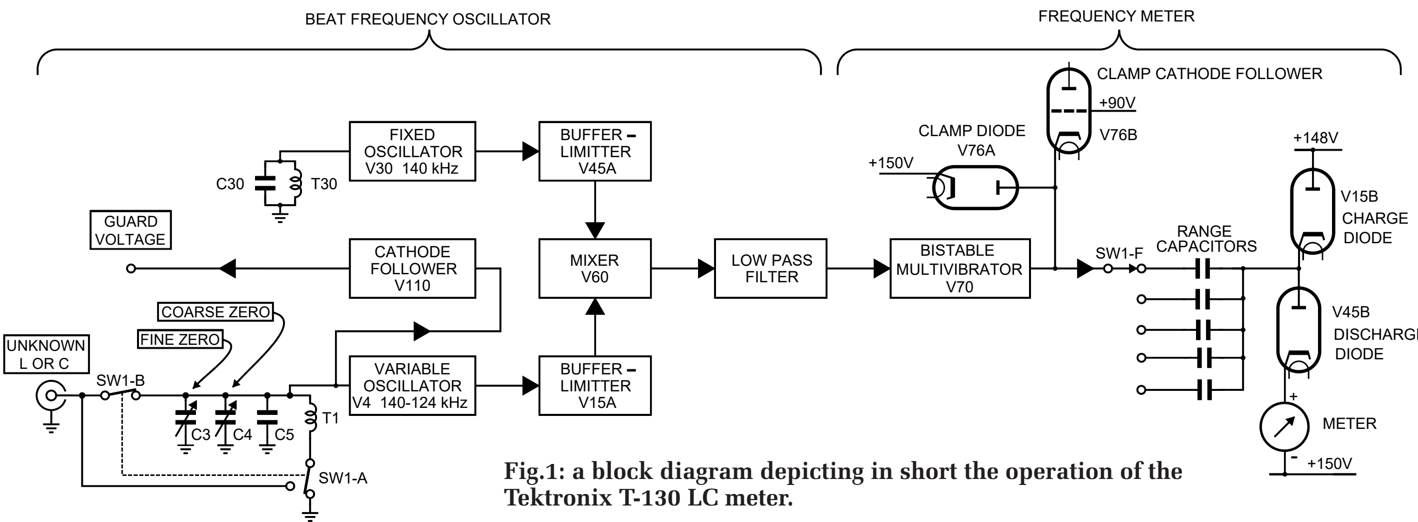


Fig.1: a block diagram depicting in short the operation of the Tektronix T-130 LC meter.

The T-130 was not intended for sale to Tek's customers – it was purely for use in the factory. It therefore wasn't designed and engineered to quite the same standards as Tek's catalog products. It was quite cramped inside, with components hidden under other parts, compromising ease of repair. But it used innovative circuitry, offered excellent performance and was easy to use.

Factory visitors noticed it in use, and many asked if they could buy one. So it was cleaned up and documented, with production beginning in 1954. It remained in the catalog until 1975, indicating just how good an instrument it was.

How it works

It operates on the beat-frequency oscillator principle. Refer to the block diagram, Fig.1; a built-in analog frequency meter responds to the difference in the frequency of two oscillators. The capacitance (or inductance) under test forms part of the tuned circuit of one of the oscillators, thus shifting its frequency.

The fixed oscillator runs at 140kHz, set by tuned circuit C30/T30.

With RANGE SELECTOR switch SW1 in any of the "µµF" (picofarad) positions, the variable oscillator is tuned by T1 and the capacitance connected to the UNKNOWN jack plus capacitors C2-C5. With SW1 in any of the "µH" positions, the tuned circuit comprises C3-C5 and T1 in series with any inductance connected to the UNKNOWN jack.

C3 and C4 are adjusted to get 140kHz from the variable oscillator with whatever wiring or cabling capacitance or inductance appears on the UNKNOWN jack. When the capacitor or inductor under test is connected, the variable oscillator frequency drops below 140kHz in approximate proportion to its value.

An LC oscillator's frequency is proportional to the square root of total tuning capacitance and to the square root of total inductance; but in this case, the change is kept approximately linear by keeping the highest calibrated inductance or capacitance under test to a small fraction of the total. The meter scales are calibrated to match.

After passing through buffers (operating in an overdriven, limiting mode) to prevent the oscillators from coupling together and synchronising, the two frequencies are mixed, and a low pass filter substantially removes all but the difference frequency. The difference frequency is approximately 62Hz per UNKNOWN pF or µH, and is fed to a bistable circuit (Schmitt trigger) to make the waveform rectangular.

Each time the multivibrator output jumps to its low level, the 'clamp cathode follower' turns on and holds the output very close to +90V (set by 100kΩ resistor R78), as the impedance of a cathode follower is 1/gm – in this case, 160Ω. The selected range capacitor is charged to +150V less the 90V via the charge diode. The amount of charge is always the same.

Each time the multivibrator output jumps to its high level, the cathode follower is cut off, and the clamp diode limits the voltage to very close to +150V. The range capacitor is discharged via the discharge diode into the meter. The meter thus receives a pulsating direct current with an average magnitude accurately proportional to frequency.

The history of Tektronix

Tektronix was founded in December 1945 by four friends: Howard Vollum, a young engineer/physicist; Jack Murdoch, radio technician; Glen McDowell, accountant; and Miles Tippery, who served with Murdoch and McDowell in the US Coast Guard during World War II. Vollum was the president and chief engineer.

Tektronix, or "Tek" as it became known, started at the beginning of the post-war golden age of the American electronics industry. Their innovative and high-class products led to rapid growth.

This was a time when the captains of industry were often engineers, passionate about making the very best of products. This includes the founders of HP, Bill Hewlett and Dave Packard, the Varian brothers with Hansen and Grinzton at Varian Associates, Melville Eastham at General Radio and Howard Vollum, passionate about oscilloscopes, at Tek.

It was quite different from today's business leaders, who seem to care much more about the financial side of the business than the 'nitty-gritty'.

Tek focused on laboratory-quality oscilloscopes and quickly revolutionised the industry, driving the US oscilloscope leader DuMont out of the market.

Why 140kHz?

As readings go below about 0.3pF (difference frequencies <18Hz), the meter pointer increasingly shakes, as the pointer then responds to individual pulses from the multivibrator. So you wouldn't want the oscillator frequencies to be any lower.

Resonance at 140kHz occurs with values of L and C of 1136μH and 1136pF respectively. These values are sufficiently larger than the instrument's top range of 300μH and 300pF full-scale that the meter is acceptably linear. You wouldn't want it any less linear.

When the instrument was designed (about 1951), very few electronics laboratories had a frequency counter, so some other method was needed for calibration. While folk involved with radio transmitters had analog heterodyne frequency meters such as the BC-221, everybody had an AM radio receiver.

In most parts of North America, high-power clear channel broadcast stations were easily received at frequencies that were multiples of 140kHz, such as WLW (700kHz), WHAS (840kHz) or KMOX (1120kHz).

So, by running a wire from the buffer output to near the radio antenna, you could tune for a null beat note, and thereby set the fixed oscillator very accurately.

And if you could not pick up a clear channel station, you could probably receive a local station on 980kHz – the 7th harmonic of 140kHz. If you couldn't do that, the 5th harmonic from the T-130 could be nulled against the 7th harmonic from your trusty 100kHz quartz reference oscillator.

Careful and thoughtful design

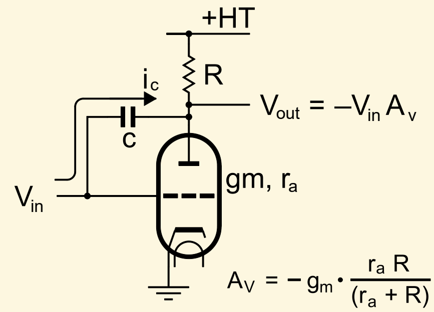
The full circuit is shown in Fig.2; it's quite complex for an LC meter. But it's clear that Cliff Moulton took care with the design to ensure the instrument is stable and accurate.

Many cheap capacitance meters employ the capacitor under test as the timing element in a multivibrator, and so interpret high leakage or shunt resistance as increased capacitance. But the T-130 substantially ignores resistance unless it lowers the Q enough to stop oscillation.

So the instrument either reads correctly or not at all. This is explained further in the panel detailing the oscillator design.

The Miller effect

The Miller effect is where any capacitance between the input and output of an inverting amplifying stage (triode, pentode, transistor, FET, op amp etc) makes the input impedance appear to include a much larger shunt capacitance.



In the circuit shown, Vout appears across the load R in parallel with the valve internal anode resistance r_a. The stage voltage gain for low values of C (ie, where the reactance of C is much larger than R) is A_V = -gm × r_a × R ÷ (r_a + R). The negative sign denotes phase inversion.

For typical triodes in typical circuits, Av is around -10 to -40. The capacitor then sees a voltage across it of ($V_{IN} + A_V \times V_{IN}$), ie, $V_{IN} \times (1 + A_V)$, and its current is thus increased by the A_V term.

Since the capacitor current is also included in the input current, the input impedance (the load on the previous stage) appears to include, in addition to the grid-cathode capacitance, a shunt capacitance of C × (1 + A_V) or approximately 10-40 times C.

The capacitor C comprises tube internal grid-anode capacitance, tube socket capacitance and any stray capacitance due to proximity of grid wiring to anode wiring.

The Miller effect with triodes, by its large capacitive load on any previous stage, typically causes the bandwidth of the preceding stage to be a small fraction of what it otherwise would be.

For more details, see John M Miller, Dependence of the input impedance of a three-electrode vacuum tube upon the load in the plate circuit, Scientific Papers of the Bureau of Standards, 15(351), pp367-385, 1920, USA.

A close-up of part of the variable oscillator section, incorporating V4 and variable capacitors C2-C5, as described in the panel labelled "An ingenious oscillator design".

The cathode interface layer

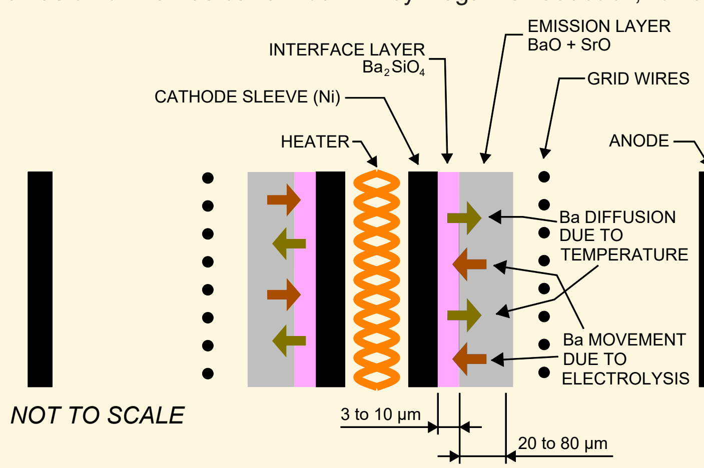
The nickel used in cathode sleeves before the early 1950s usually contained trace amounts (~0.05%) of silicon. During factory processing, and sometimes during early service, silicon diffuses to the surface and reacts with barium oxide. This forms a microscopically thin 'interface layer' of barium orthosilicate between the nickel sleeve and the oxide emission layer:

$$Si + 4BaO \rightarrow Ba_2SiO_4 + 2Ba$$

Pure barium orthosilicate has very high resistivity. As the interface layer is so thin and has free barium atoms within it, the resistance is low, and it does not initially affect tube operation. During tube operation, the high temperature required for emission drives diffusion of the free barium out of the interface layer, increasing the resistance.

Fortunately, cathode current causes barium atoms to diffuse back into the interface layer via an electrolysis process. The balance of these opposing effects results in interface resistance being quite sensitive to heater voltage. A 10% drop in heater voltage reduces cathode temperature by about 3.5% and interface resistance for a given cathode current by about 50%. The diffusion processes are very slow.

Interface layer resistance has the same effect as any resistance in series with the cathode; it increases cathode bias, possibly biasing the tube back to where the gain is lower, and also, by negative feedback, lowering gm.



Note that although the tube may test low for gm, its emission can be entirely normal.

A tube with low gm due to the interface layer can usually be rejuvenated by operating it in a tube tester or rejuvenator with the maximum rated cathode current for a few days or more. This is not to be confused with rejuvenating a low emission tube by running it with a high heater voltage, which often doesn't work. And if it does, it's only for a while.

As the interface layer is so thin, it makes a pretty good RF bypass capacitor for its own resistance. Thus, you can easily detect the presence of an interface layer by measuring gm at an audio frequency and at RF, say 2MHz. The gm at 2MHz will be normal (unless the valve has some other fault), but the gm at audio frequencies will be lower.

Valves manufactured after about 1955 generally have high-purity cathode sleeves (less than 0.001% silicon), markedly reducing interface layer thickness and avoiding these problems.

Reference: M. R. Child, The Growth and Properties of Cathode Interface Layers in Receiving Valves, The Post Office Electrical Engineers' Journal, Vol 44[4], pp176-178, London 1952.

The variable oscillator operates under starvation conditions – very low anode and screen current – which results in a high gain. This means only 600mV peak-to-peak on the tuned circuit, even though the output to the buffer is quite high.

The low amplitude on the tuned circuit not only reduces the chance of forward-biasing junctions when in-circuit testing. It also means that the T-130 can be used to measure the Miller effect, as typical triode circuits under test will not be driven into overload. If you aren't familiar with the Miller effect, see the panel with the same name at upper left.

Running a valve under starvation conditions gives a high space charge density. The 6U8 triode-pentode variable oscillator valve (V4) has its heater voltage reduced by 1.5Ω resistor R405. This reduces the effect of any interface layer and reduces space charge, so oscillator drift with AC mains voltage better matches the fixed oscillator.

See the panel later in this article for an explanation of space charge density, and above for the interface layer.

The meter is pegged to the +150V rail and not ground as might be expected. This reduces the average DC voltage across the range capacitor, so that it's much less likely to develop leakage, and any leakage won't matter as much.

Bistable multivibrator

The circuit around V70 is called a bistable multivibrator by Tektronix but will be known to most people as a Schmitt trigger, after American Otto H. Schmitt, who invented it in 1934. Considerable positive feedback via common-cathode 5.6kΩ resistor R71 forces the pentode section, V70A, to operate in two fixed states – cut off, or drawing 4.2mA anode current.

When triode V70B is cut off, pentode V70A is on, due to the voltage divider R73 and R72 (470kΩ & 180kΩ respectively). 43V is dropped across R71 – a pentode cathode current of 7.7mA. Hence, the screen-to-cathode voltage is 110V, and the 6U8 data sheet shows that the screen draws 3.5mA at this voltage. Hence the anode current is 4.2mA (7.7mA - 3.5mA).

When the input from the filter rises above V70B's grid cut-off level (about 37V), V70B begins to turn on, reducing the voltage to V70A's grid. So V70A begins to turn off, dropping the voltage on R71. This turns on V70B harder, and the circuit immediately snaps over to V70B fully on with V70A cut off.

C73 compensates for wiring and socket stray capacities and ensures the snap action is fast.

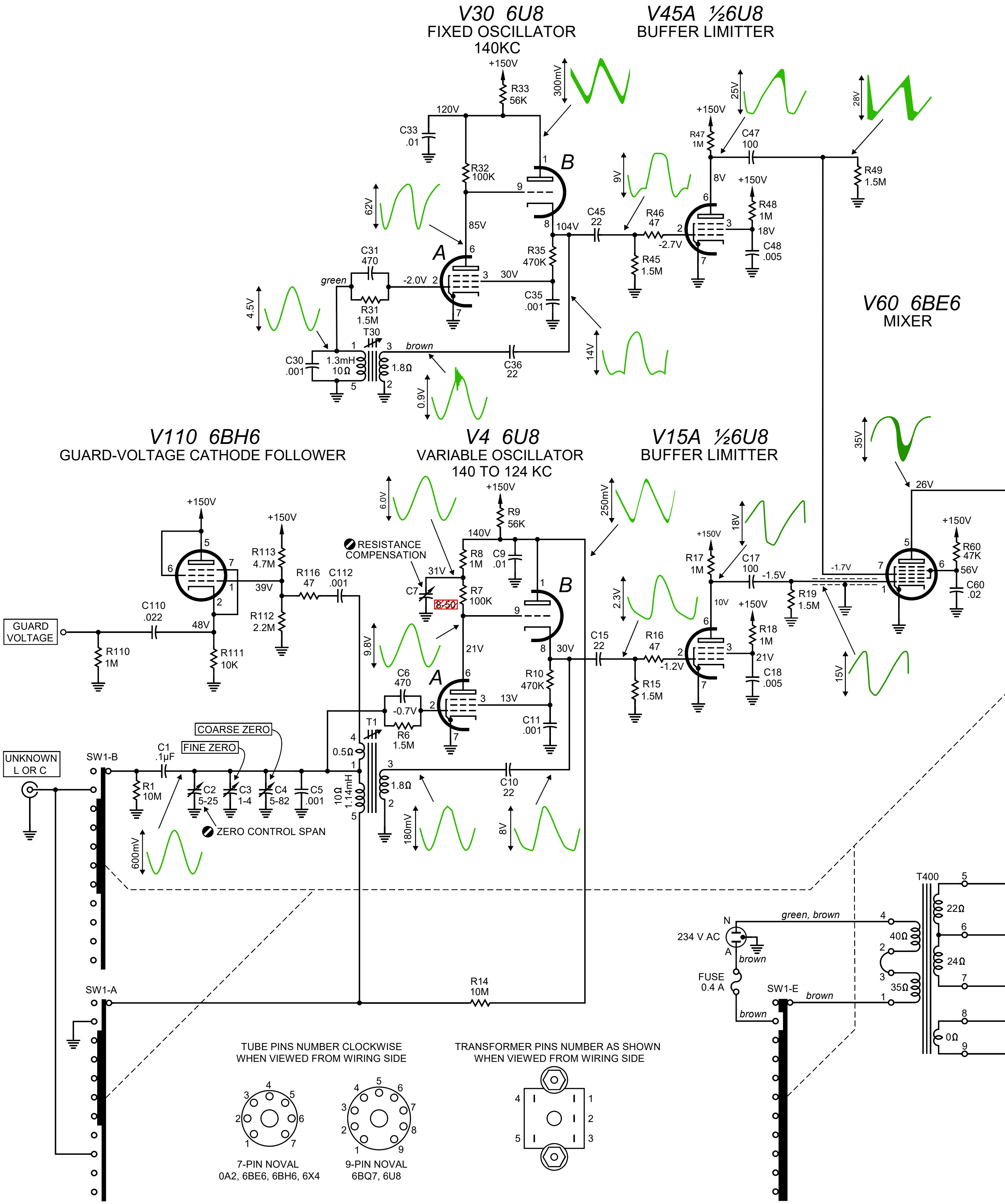
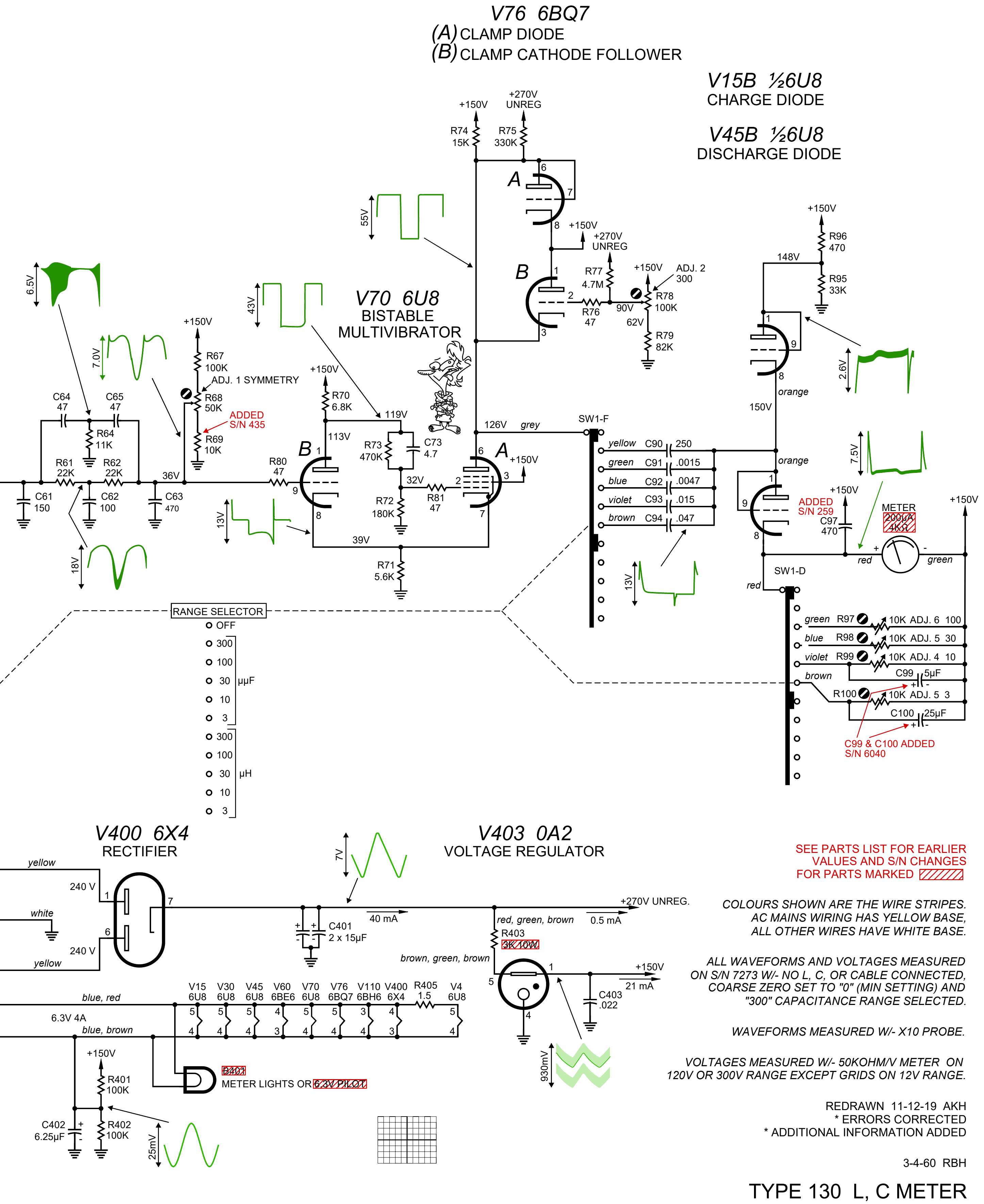


Fig.2: complete circuit diagram for the Tektronix T-130 LC meter.



V76 6BQ7
(A) CLAMP DIODE
(B) CLAMP CATHODE FOLLOWER
V15B ½6U8
CHARGE DIODE
V45B ½6U8
DISCHARGE DIODE
V70 6U8
BISTABLE
MULTIVIBRATOR
ADJ. 1 SYMMETRY
ADDED S/N 435
ADJ. 2 300
ADDED S/N 259
METER
SW1-F
SW1-D
RANGE SELECTOR
OFF
300
100
30
10
3
µµF
300
100
30
10
3
µH
C99 & C100 ADDED S/N 6040
V400 6X4
RECTIFIER
V403 0A2
VOLTAGE REGULATOR
+270V UNREG.
C401 2 x 15µF
METER LIGHTS OR 6.3V PILOT
SEE PARTS LIST FOR EARLIER VALUES AND S/N CHANGES FOR PARTS MARKED
COLOURS SHOWN ARE THE WIRE STRIPES.
AC MAINS WIRING HAS YELLOW BASE,
ALL OTHER WIRES HAVE WHITE BASE.
ALL WAVEFORMS AND VOLTAGES MEASURED
ON S/N 7273 W/- NO L, C, OR CABLE CONNECTED,
COARSE ZERO SET TO "0" (MIN SETTING) AND
"300" CAPACITANCE RANGE SELECTED.
WAVEFORMS MEASURED W/- X10 PROBE.
VOLTAGES MEASURED W/- 50KOHM/V METER ON
120V OR 300V RANGE EXCEPT GRIDS ON 12V RANGE.
REDRAWN 11-12-19 AKH
* ERRORS CORRECTED
* ADDITIONAL INFORMATION ADDED
3-4-60 RBH
TYPE 130 L, C METER

Shown above is the T-130 testing an MSA 100pF capacitor, which returned a reading of ~98pF. Below is a short description of the controls on the front panel:
<u>RANGE SELECTOR</u>: an 11-position switch (five each for capacitance and inductance), which also functions as the power switch.
<u>COARSE ZERO</u>: used to adjust for capacitance in connecting leads or connectors.
<u>FINE ZERO</u>: finer range adjustment compared to COARSE ZERO.
<u>GUARD VOLTAGE</u>: used to cancel out the influence of any other component connected to the part under test.

While V70B is on, it acts as a cathode-follower and thus the voltage across R71 is about 2V more than the input voltage at V70B's grid. When the input from the filter is reversing later in the cycle and drops to about 35V, V70B starts to turn off, turning on V70A via the voltage divider formed by R72 and R73. V70A then raises the voltage across R71, forcing V70B further off and the circuit snaps back.

Thus, V70A snaps from cut-off to drawing a constant 4.2mA when the filter output rises above 37V, and snaps back to full cut-off when the filter output falls below 35V. The filter output considerably exceeds this range.

Guard cathode follower

This socket connects to the RANGE SELECTOR on the front panel. The visible ring connects to V70's anode, and the crimped lugs of the ring on the other side connect to the 230V AC mains input.

V110 (6BH6) is a pentode connected to function as a triode cathode-follower. It takes a signal from the variable oscillator tuning coil and makes it available as a low-impedance (250Ω) guard signal on the front panel.

Since the voltage gain of a cathode follower is slightly less than unity, the cathode follower is driven from an over-wind on the tuning coil to compensate.

You can connect the guard output to the other end of any components connected to the item under test. Because there is then the same voltage at both ends of these components, the T-130 ignores them and gives a true reading.

Power supply

V400, a 6X4, rectifies the AC from the power transformer to derive the unregulated 270V HT rail. A 0A2 (V403) regulates the 150V rail. The 0A2 is a cold-cathode gas-filled valve that performs the same function as a zener diode.

The valve heaters are run at 75V above ground. This is because the heater-cathode rating of the valves is only 100V. Since some cathodes are at or near ground, and some are at +150V, the heaters are run halfway between to keep all valves within their ratings.

Next month

That concludes the description of how the T-130 works. But what about the one that I purchased? What was wrong with it? How did I fix it? Don't worry; I have documented all the work in detail.

It will be described over the next two issues, starting with the aesthetic restoration and finishing up with circuit repairs and calibration.

Space charge capacitance

Valve cathodes are typically designed to emit electrons at about 2.5 times the rated maximum cathode current.

Taking the 6U8 pentode as an example, the rated maximum cathode current is 13mA, so the emission should be 33mA. In typical use, the sum of the anode and screen current would be around 4mA due to negative grid bias. The current is even less in the T-130 variable oscillator valve (V4).

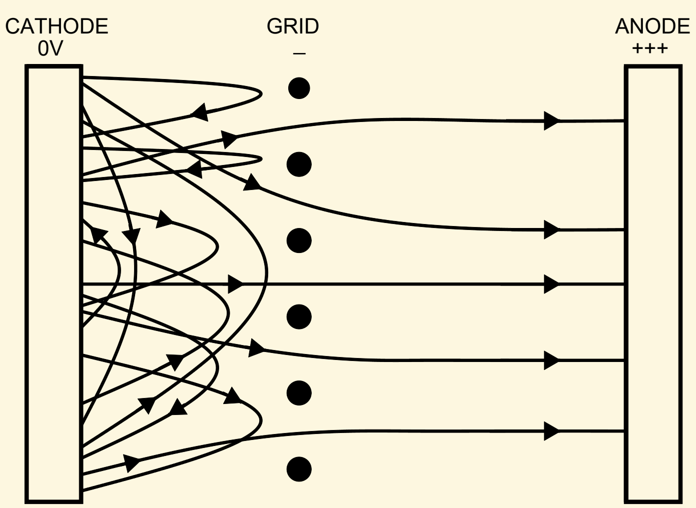
So if the cathode is emitting 33mA, and only 4mA is getting past the grid, what happens to the remaining 29mA? It goes back into the cathode!

In any conductor, conduction electrons are in continuous motion whizzing about at random velocity and direction. Collisions with atoms continually cause electrons to change direction. But at ordinary temperatures, practically none have enough inertia to escape the conductor due to the attraction of nearby nuclei – if electrons are not bound to particular nuclei, the nuclei must have a positive charge.

By heating the cathode, we raise the velocity of the conduction electrons so that some have enough inertia to escape. Any electrons leaving the cathode that are more than the number required to make up the anode current (which must return to the cathode via the external circuit) leave a positive charge in the cathode. So these excess electrons are inevitably sucked back into the cathode.

They follow individual parabolic paths outside the cathode, much like stones thrown up into the air returning to the ground. Negative grid bias encourages more of these electrons to give up and return to the cathode.

The cloud of electrons between the cathode and grid is called a "space charge" and tends to self-limit in local density, as space charge electrons repel more electrons leaving the cathode.

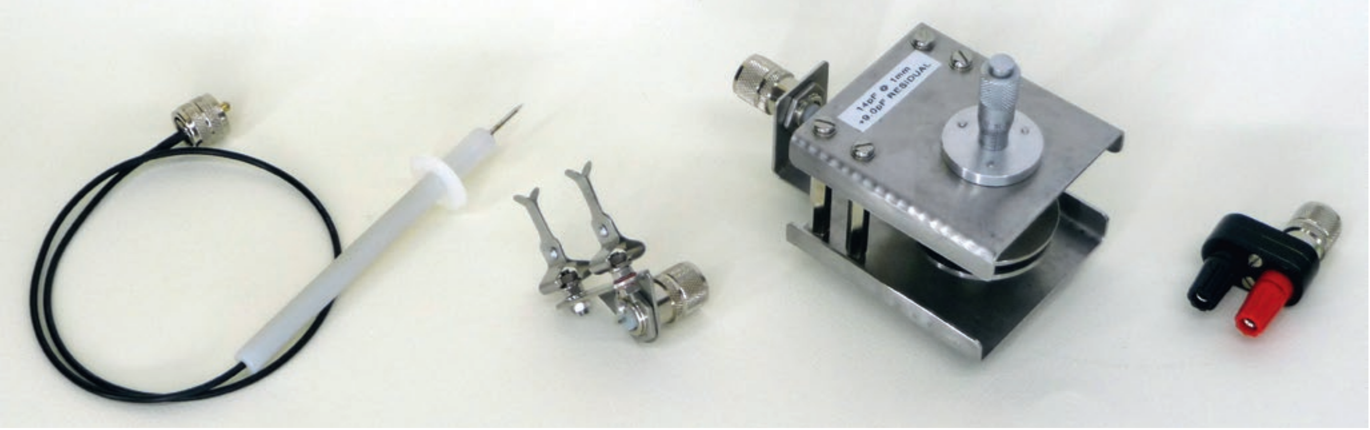


But it is considerably denser than the electron density between the grid and anode. The lower the anode and screen current, the denser the space charge. Our 6U8 example cathode always emits 33mA, but it may have up to 33mA returning.

The space charge electrons are in frequent contact with the cathode, and can be influenced by a varying electric field, so they constitute an electrical conductor, just as electrons do within a metallic conductor. So, we have a conductor – the space charge – near to, but not touching, another conductor – the negative grid. That's a capacitor! And it has a plate spacing less than the physical grid-cathode spacing.

The space charge capacitance typically adds 0.5-2.5pF to the inherent capacitance of the grid-cathode structure. This capacitance decreases with increasing grid bias (a more negative grid pushes the space charge further back toward the cathode) and increases with decreasing anode + screen current.

It increases about 10% for each 1% increase in heater voltage; hence, heater voltage variation due to AC mains variation is a significant cause of frequency drift in grid-tuned oscillators. An increase in heater voltage causes a decrease in oscillator frequency.



Shown above are a variety of homemade adaptors which can be connected to the UNKNOWN jack on the front panel. The largest one (second from the right) is a variable space capacitor for measuring permittivity – the degree that an insulating material increases capacitance between the plates over the capacitance obtained with air or vacuum spacing.

The fundamental requirements of a sinewave oscillator are:

- Something to set the frequency – a tuned circuit
- An amplifier to make up for the inevitable losses in the tuned circuit by feeding some of its output back to the tuned circuit – "tickling" the tuned circuit
- Feedback in-phase with the tuned circuit oscillation.
- A means to control the oscillation level

Often the amplifier was a single grounded-cathode valve that inverts the phase. This is corrected by connecting the tickler winding to give a second phase inversion.

Figure A shows a typical AM radio oscillator at mid-band. Let's take a look at how it works, and how the T-130 oscillators differ.

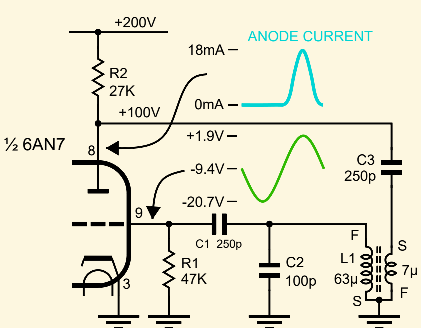


Figure A: a typical AM radio oscillator configuration. The T-130's implementation is shown at lower right in Figure D.

100pF capacitor C2 (comprising one section of the gang, a trimmer, and padder if used) and inductor L1 form the tuned circuit. The optimum oscillation voltage on the grid is 8V RMS, ie, 23V peak-to-peak. Grid current flows briefly on the positive peaks, clamping the tip of the peaks to about +1.9V. This forces the average grid voltage to be -9.4V by charging C1.

The 6AN7 triode section has a semi-remote cut-off, beginning at about -3V and fully cut off at -10V. Thus, significant anode current flows for only about 120° – as shown in Figure B.

250pF capacitor C3 and the tickler winding offer a low impedance, so almost all of the AC part of the anode current flows in the tickler winding, and only the DC part, about 3.8mA, flows in 27kΩ resistor R2. The valve works quite hard, conducting 18mA peak.

Oscillation always starts because the anode current without oscillation (and so no grid bias) is 5.1mA and gm (transconductance) is maximum at this level – as shown in Figure C.

The oscillation amplitude is regulated because if the grid oscillation increases, a greater fraction of the sinewave is beyond cut-off. As the grid will not allow any increase in the positive direction, the peak anode current is fixed at about 18mA. Still, the grid excursion goes further beyond cut-off, so the valve conduction angle decreases.

Therefore, the energy fed back via the tickler winding decreases, holding back the increase at the grid. This is called grid-controlled amplitude or grid stabilisation. Almost all LC valve oscillators use grid stabilisation.

R1 is typically 47kΩ. A much higher value is not used as it will let the circuit 'squeg', ie, multivibrate at a lower frequency and amplitude modulate the desired oscillation. R1 dissipates 1.36mW due to the AC component of the waveform, and a further 1.88mW due to the DC average voltage. 0.38mW is lost in grid dissipation.

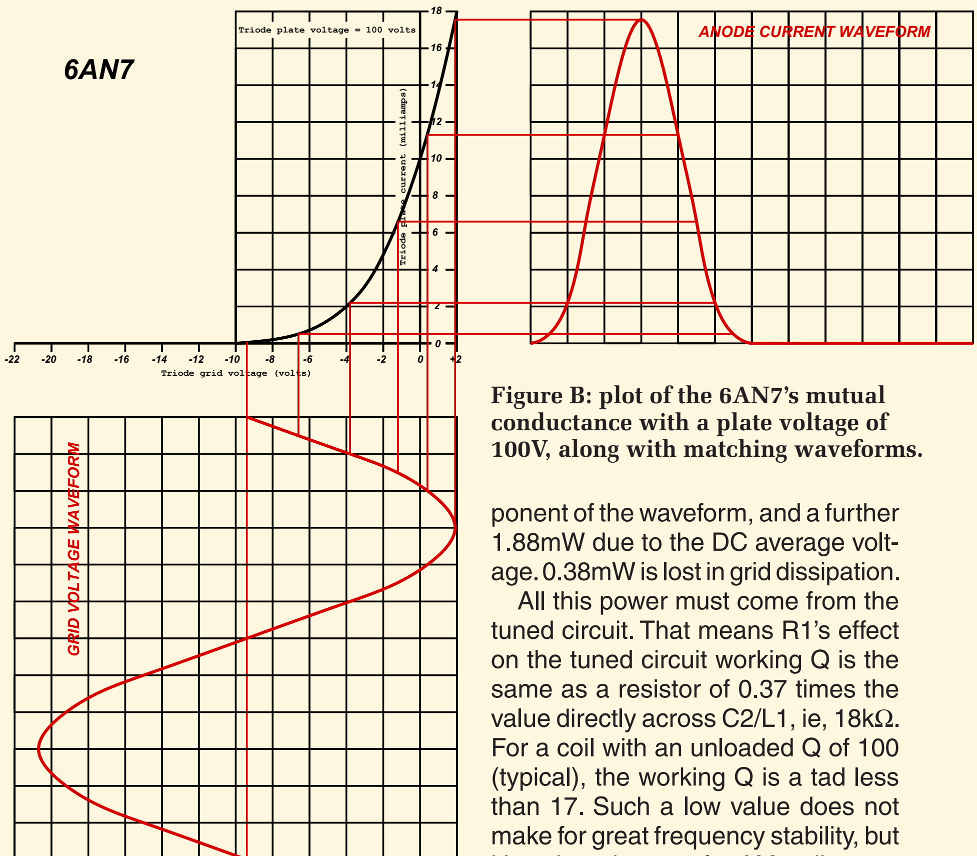


Figure B: plot of the 6AN7's mutual conductance with a plate voltage of 100V, along with matching waveforms.

All this power must come from the tuned circuit. That means R1's effect on the tuned circuit working Q is the same as a resistor of 0.37 times the value directly across C2/L1, ie, 18kΩ. For a coil with an unloaded Q of 100 (typical), the working Q is a tad less than 17. Such a low value does not make for great frequency stability, but it's quite adequate for AM radio.

Figure D shows the T-130 Variable Oscillator. The fixed oscillator is identical except for its operating level.

The pentode stage operates as a Class-A voltage amplifier under starvation conditions. This provides a high output level with only 0.3V peak on the tuned circuit, comprising C2-C5 and T1. This low level is essential for in-circuit testing, especially when using the T-130 to measure Miller effect capacitance.

The pentode is biased not by grid rectification but by its own space charge. The grid never goes positive and never draws energy from the tuned circuit. Since the energy dissipated in 1.5MΩ resistor R6 comes from the pentode space charge and not from the tuned circuit, the tuned circuit operates at its unloaded Q.

Since the grid never goes positive and doesn't rectify, the circuit cannot squeg no matter how high the grid resistor (R6) is.

For an iron dust core of the size used, the Q is probably about 150-200. It will be lowered by resistance in the circuit under test, of course, but

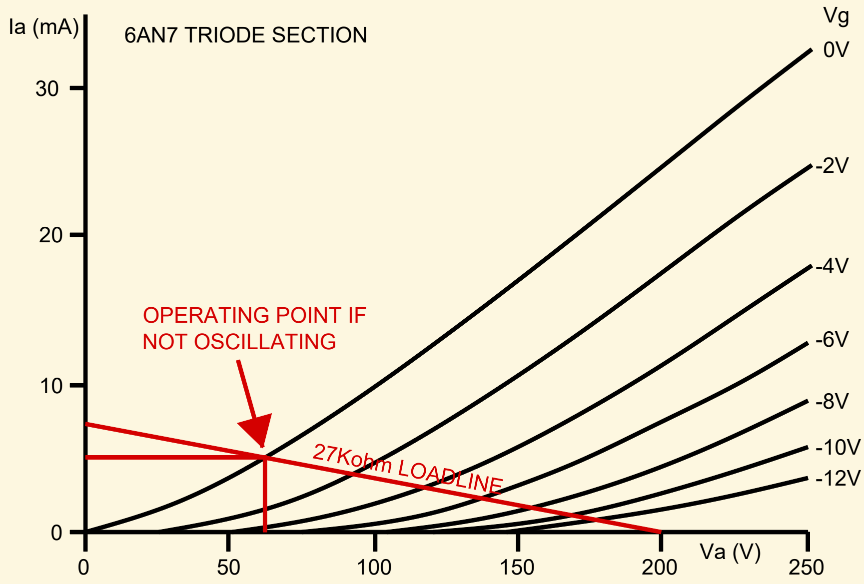


Figure C: plot of the 6AN7's anode voltage versus anode current for various grid bias amounts. The 27kΩ is the load connected to the plate of the 6AN7 (R2 in Figure A).

will always be above 30, and usually well above. The low-impedance tickler winding is loosely coupled and 'looks into' a small capacitance (22pF capacitor C10). So the tickler has no significant effect on Q.

The triode only conducts on positive peaks, as C10 can be charged by the cathode but not discharged by it. The triode conducts for only about 80°. That's why the signal at the cathode is half what it is at the grid. The cathode current peaks at 280µA; during the peaks, 120µA flows in C10, 120µA in C15, and 36µA in R19. The pentode current averages 110µA. The 6U8 is far from being worked hard.

If the oscillation level increases, C10 and C15 will charge up a bit more so that the signal on 470kΩ resistor R10 remains at about 6V peak-to-peak. But the greater swing on the grid means that the triode conduction angle must decrease. So less energy is fed back to the tuned circuit.

Unlike most LC tuned oscillators, this circuit is cathode-regulated. By using a triode-pentode with cathode stabilisation, we get a very stable oscillator. Considerable negative DC feedback via R10 holds the DC working point close to the designed level regardless of valve aging.

Ideally, signal feedback in an oscillator should be in-phase. What happens if it is not precisely in-phase? The first effect is that slightly more amplifier gain is needed. That's unimportant; plenty of gain is available, and the circuit will self-adjust anyway. The second effect is important in this application: it changes the frequency slightly.

Say the feedback is slightly late. By holding back the rate of change in the tuned circuit, the frequency drops slightly. Conversely, if the feedback is a little early, the rate of change is reinforced, and the frequency increases.

The ordinarily high Q of the tuned circuit strongly resists this influence over frequency. This means that if Q is lowered, say by a resistance across the tuned circuit, the frequency will change in the direction pulled by the feedback phase.

The pentode output is phase-inverted and of high impedance; about 800kΩ. Variable capacitor C7, together with stray wiring capacitance and the grid-anode capacitance of the triode section (~2pF), causes an additional phase lag of about 80°. So the signal at the triode grid, and the cathode, is lagging by 260°.

Most of the triode output voltage is dropped across C10, which means that C10 causes a phase lead, of about 80°. So we are back to approximately 180°, and, like many oscillator circuits, the situation is corrected by the phasing of the tickler winding (between pins 2 & 3 of T1).

Part of the calibration procedure is to adjust the phase by adjusting C7 so that the frequency doesn't change when two different test resistances are connected across the UNKNOWN terminals. This means that the feedback is precisely in-phase, and the T-130 reading is independent of any shunt resistance when in-circuit testing – within reason. Clever, eh? Too much loss stops oscillation.

Correct adjustment of C7 also means that the variable oscillator is maximally tolerant of contact resistance in the RANGE SELECTOR switch, improving frequency resetability. **SC**

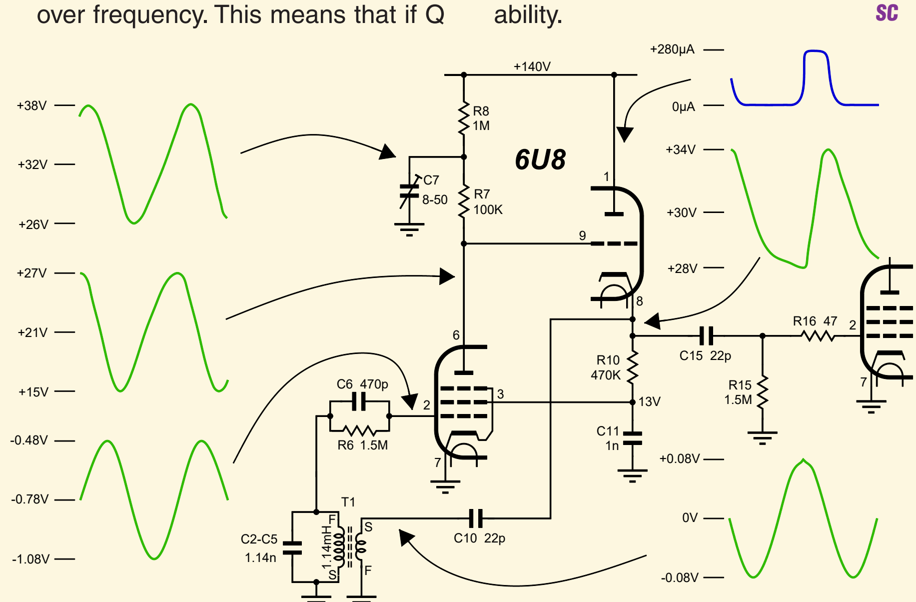


Figure D: the variable oscillator configuration used in the T-130 uses a 6U8 triode pentode.