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VR-005B

Vintage Radio Application Note

UPGRADING THE HEATHKIT MODEL TT-1/TT-1A TUBE TESTER TO THE TT-1B

Rev B01/07-25/vr_apl_005b
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VR-005A/B FRONT-MATTER

(*IMPORTANT - READ THESE PAGES FIRST*)

VR-005B:

VR-005B is an upgrade to VR-005A that replaces the original command line input Gm computation software with a modernized and improved graphical user interface version. This new software is presented in Appendix H.

VR-005A:

I ERRATA LIST

VR-005A contains important updates to the original paper (VR-005) that correct errors in both the schematics and text. The complete errata list is as follows:

A. FIGURE 8 SCHEMATIC (Pg. 18)

1. The wattage rating of R14 (270 ohms) was upped from 1/4W to 1/2W for extra margin.

2. The LCA110 opto-isolator switch had no reference designator even though it was referred to as U1 in the text. The schematic was amended to add this U1 reference designator.

3. IC numbers U2-U6 as referred to in the text were incorrectly labeled on the schematic as U1-U5, respectively. The schematic was amended so that these ICs are now labeled U2-U6 in conformance with the text.

4. The U2 (PIC 16F628A microprocessor) pin 5 ground was incorrectly shown as a chassis ground symbol rather than as a local ground symbol. The schematic was amended to correct this. (For safety reasons, Heathkit designed the TT-1/TT-1A so that the chassis was isolated from all the electronics.)

5. C8 was incorrectly designated as a .1u/MYL capacitor. Its actual value is .01u/MYL. (The MYL suffix designates a temperature-stable Mylar capacitor.) The schematic was amended to correct this.

6. Duplicate connectors were present for J2:5, J2:6, P2:5, and P2:6. These duplicates were removed.

7. Miscellaneous non-circuitry changes were made to enhance clarity and improve presentation.

B. FIGURE 11 SCHEMATIC (Pg. 22)

1. The schematic was amended to correct some wiring errors.

C. FIGURE 19 SCHEMATIC (Pg. 33)

1. Capacitors C1-C5 and C7 were not labeled as Mylar capacitors. Since these capacitors should be temperature-stable, the schematic was amended to label them as Mylars (MYL).

2. The Relay+ pin was incorrectly labeled as P2:4 rather than P4:4. The schematic was amended to correct this.

3. Miscellaneous non-circuitry changes were made to enhance clarity and improve presentation.

D. TEXT ERROR (Pg. 22)

1. The text was amended to conform with the Figure 11 wiring error corrections (see above).

E. TEXT ERROR (Pg. 32)

1. The meter terminal pin designations were listed incorrectly in the first sentence of the last paragraph as follows (emphasis added):

   "Note that this meter drive voltage is applied to the TT-1A DC meter terminals (pins ***1*** and ***2***) rather than its AC meter terminals (pins ***3*** and ***4*** as per the original design)."

2. This sentence was therefore corrected as follows (emphasis again added):

   "Note that this meter drive voltage is applied to the TT-1A DC meter terminals (pins ***3*** and ***4***) rather than its AC meter terminals (pins ***1*** and ***2*** as per the original design)."

II SOFTWARE ENHANCEMENT

The computer program (TT-1B_01.BAS) used to determine the various TT-1B switch settings and compute the appropriate accept/reject meter indication and other information requires an MS-DOS PC. A new version of this software that runs on modern Windows PCs has been developed. See Appendix E for a fuller explanation.

III UPCOMING ENHANCEMENTS FOR THE TT-1B

Important enhancements to the TT-1B are underway as of this writing that will further modernize this instrument. A preview is provided in the added Appendix G.

IV ACKNOWLEDGMENT

All of the errata items listed above were found by Mr. Steve Mazur of Mazur Sound in Austin, Texas. Steve discovered and reported these errors while successfully converting his TT-1 to a TT-1B as per the original VR-005 paper.

Although tube testers are fairly straightforward conceptually, they are quite complicated in practice due to the large number of switches and sockets and their associated complex wiring. This issue is compounded by the impracticality of drafting a compact schematic based on a progressive signal flow.

As a result, an effort as complicated as upgrading the TT-1A to the TT-1B is not only difficult to implement, but also difficult to document. The real test of the quality of the documentation is the ability of the reader to use it to successfully implement the upgrade. A great deal of skill and effort was required on Steve's part to find these documentation errors and successfully implement the TT-1B upgrade to his unmodified TT-1.

Steve is now a contributor to this project and is working with this author on the upcoming enhancements to the TT-1B as per the added Appendix G.

We similarly hope that other readers embarking on this project will report any errors that they may find, and offer any suggestions they may have to improve the clarity of this paper.

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LIST OF ILLUSTRATIONS

SECTION I - INTRODUCTION

In my previous paper (VR-004), I discussed the various issues associated with selecting, procuring, reconditioning, and troubleshooting the Heathkit TT-1/TT-1A transconductance tube tester. The greater purpose of VR-004 was to present detailed information for the benefit of readers wanting to restore the instrument to "like-new" condition and to have the instrument fully meet its original factory specifications. I strongly recommend that readers first read VR-004 before reading this paper.

To briefly summarize VR-004, I purchased two TT-1A tube testers on Ebay. The first was a "primo" unit that was very clean and in excellent mechanical condition but did not function properly due to assembly errors and faulty components. The second was a lower-priced "junker" unit that was shop-worn and dilapidated, but wired correctly and mostly functional. With the benefit of having both of these units for comparison, I was successful in my efforts to recondition both units.

Although both TT-1As passed the various test and calibration procedures presented in the Heathkit TT-1/TT-1A assembly manual, I found that they indicated significantly different values of transconductance for the same 6AU6 tube. In Section V (Closing Comments) of VR-004, I offered some preliminary thoughts as to why this was the case and pledged to write a second paper presenting refinements and enhancements to mitigate this problem. This paper documents those improvements and related topics. With the benefit of these improvements, TT-1/TT-1A transconductance measurement accuracy is greatly enhanced.

Since the upgrades are extensive to the point of being a major redesign effort, I have dubbed the upgraded unit the TT-1B. Whereas a reconditioned TT-1A is a capable mid-range vintage transconductance tube tester, the TT-1B upgrade presented in this paper converts the TT-1A to a fine precision instrument with unmatched measurement accuracy.

For the sake of brevity and consistency, the TT-1/TT-1A is heretofore referred to as the "TT-1A". As per VR-004, the TT-1A is actually a TT-1 with an added TTA-1-1 supplemental tube socket adaptor assembly that Heathkit added to its product line in 1962 to accommodate later vacuum tubes (including Nuvistors and 12-pin Compactrons). The TTA-1-1 was back-compatible with all previously sold TT-1s. Aside from the inclusion of the TTA-1-1, there are no significant differences between the TT-1 and the TT-1A.

Be sure to visit my N6DC vintage radio website for possible revisions to this paper as well as other vintage radio technical articles at www.rdfproducts.com/N6DC.Vintage.Radio.htm.

** DANGER **

The Heathkit TT-1/TT-1A tube tester contains dangerous high voltages that can be ***lethal if contacted****. Those intending to work on the TT-1/TT-1A should be well versed in working on equipment with high voltages present and be completely familiar with all necessary safety precautions. Those unfamiliar with these safety precautions or inexperienced in working on equipment containing high voltages should not remove the TT-1/TT-1A chassis from its case.*

SECTION II - MODIFICATION SUMMARY

The primary impetus for writing this paper was the fact that the two TT-1A tube testers I reconditioned as per VR-004 measured significantly different transconductance values for the same 6AU6 tube. Also, Kent Nickerson (who wrote an excellent paper describing the steps he had taken to recondition his own TT-1A) advised me that the TT-1A transconductance measurement accuracy is likely no better than 30%.

My initial thought was that an instrument as elaborately designed as the TT-1A should have better accuracy. It was also my thought, as expressed in VR-004, that this loose accuracy might be caused by anomalies associated with plate choke (L1) impedance and the fact that the 5 kHz signal oscillator had excessive harmonic content. Although I found these to be contributing factors, the real problem was more fundamental and was based on issues associated with the AC metering circuitry and uncertainties in the load resistance (R_L) presented to the plate circuit of the tube under test.

Unfortunately, these were found to be issues fundamental to the TT-1A design that could not be remedied by simple means and incremental changes. The only path forward on this project was to completely redesign the metering and load circuitry using modern solid-state techniques. Other enhancements included stabilizing the operating voltages and replacing the 5 kHz signal oscillator with a low-distortion version using modern solid-state circuitry. At the end of the day, this project escalated in scope, magnitude, and required time and effort far beyond my original expectations. Consequently, this unforeseen escalation is reflected in the length and detail of this paper.

Section IV presents the modifications made to stabilize the operating voltages. These modifications consist of employing solid-state voltage regulation for the grid bias voltage and then using the Set Line control exclusively to precisely set the filament voltage.

Section V presents circuitry for a modern low-distortion sinusoidal output solid-state oscillator to replace the original harmonic-rich 5 kHz signal oscillator.

Section VI presents circuitry to replace the original passive signal processing and metering circuitry with much more elaborate and precise solid-state circuitry. An added easy-to-set front-panel transconductance calibration adjustment is also presented.

These modifications are extensive and time-consuming but provide far superior and more stable transconductance measurement accuracy than the original circuitry. Be aware that this project is an “all or none” proposition in that all these modifications must be done in their entirety to yield significant measurement accuracy improvement. *Also, these modifications should be done in the order presented.*

While the primary design goal was to improve the transconductance measurement accuracy, it was also a goal that the secondary instrument functions (e.g., leakage testing, diode testing, etc.) not be impaired as a result. This goal was achieved. (Although thyratrons and magic-eye tubes wee not tested due to unavailability, the TT-1B will very likely function normally for these tubes since the modifications, if properly implemented, do not affect this part of the circuitry.)

A further design goal was to cause as little disruption as possible to existing circuitry so as to minimize difficulty with circuit tracing and troubleshooting (although this goal could be only partially met since it's difficult to fix something without changing anything). To this end, marked-up sections of the TT-1A schematic have been included to clearly illustrate changes to the original circuitry and how the added modules are integrated.

Some readers may question these modifications from an "authenticity" standpoint (i.e., that implementing these modifications is actually a modernization project that is inconsistent with a vintage electronics undertaking). This is a legitimate point, especially in light of the fact that new controls are added to the TT-1A front-panel, and a point that I would concede for my more typical vintage radio reconditioning projects.

However, I procured my TT-1As with the primary goal in mind of having a functional tube tester to be used as an essential test instrument in support of my vintage radio projects rather than as a vintage electronics undertaking in and of itself. With this being the primary goal, the extensive modernizations detailed in this paper are appropriate.

Although I procured two TT-1A tube testers (one "junker" and one "primo" unit as per VR-004), I converted only one of these (the "primo" unit) to a TT-1B. Although I reconditioned the "junker" unit so that it is fully functional and meets its original specifications, the primary purpose of this unit is to serve as a parts source for the "primo" unit as required.

As a final, cautionary note, *be very careful when implementing these modifications to protect the TT-1A meter*. It is easy to inadvertently overdrive this meter during testing and troubleshooting, possibly causing damage. Since the only way to obtain a replacement meter is to procure another TT-1A, I strongly recommend using an external substitute meter until the modifications have been fully tested and verified. Since the TT-1A basic full-scale meter movement is 890 microamperes, I recommend substituting a standard 0-1 milliampere meter for this purpose.

SECTION III - PRELIMINARIES

First, the TT-1A to be modified must be a *fully functional unit that has passed all the operational tests presented in the Heathkit TT-1A assembly manual*.

Second, I strongly recommend that the TT-1A be *reconditioned as per VR-004* before attempting these modifications.

Third, both the TT-1A assembly and operational manuals must be procured to facilitate component location and troubleshooting. This documentation (which includes B-size fold-out drawings) is available from Don Peterson at Data Professionals as per VR-004.

Fourth, a digital voltmeter (DVM) and oscilloscope will be necessary to test and troubleshoot the new circuitry.

Finally, be advised that this is not a project for beginners. This project should be attempted only by readers experienced in electronic prototyping, circuit tracing, and troubleshooting. This project also requires patience, care, and attention to detail.

Addressing the matter of component selection, I have indicated 1% metal film precision resistors in circuitry where the resistance value directly affects overall measurement accuracy. Although 5% carbon film resistors can be substituted, these are less temperature-stable and only modestly less expensive than the recommended 1% metal film resistors.

For the benefit of readers who wish to use 5% carbon film resistors, however, I have specified 1% resistors with standard EIA 5% values for easy substitution. To explain, 220k is a standard EIA 5% resistor value. Although 220k is not a standard EIA 1% resistor value (the closest standard EIA 1% value is 221k), most manufacturers now offer 1% resistors in the 5% standard EIA values for convenience.

Also, for readers who choose to use 1% resistors as recommended, feel free to substitute the closest 1% standard values if you cannot find a convenient source of 1% resistors in 5% standard values (i.e., to use the previous example, you can substitute a 221k resistor for a resistor specified at 220k).

Similarly, I have indicated temperature-stable Mylar capacitors in critical circuitry (active filters and frequency-sensitive AC coupling circuitry). Do not substitute less temperature-stable ceramic capacitors for these circuits since this will result in calibration drift. Ceramic capacitors are fully acceptable, however, for all non-critical AC coupling and bypass applications.

Finally, the terms “transconductance” and “Gm” are used in this paper interchangeably.

SECTION IV - STABILIZING THE TT-1A OPERATING VOLTAGES

A. OVERVIEW

The TT-1A is required to provide filament, grid bias, and plate/screen voltages to the tube under test. The problem is that no voltage regulation is employed. This in turn requires that the Set Line control (a power potentiometer that varies the power transformer primary winding voltage) be manually adjusted for the correct Line Check meter indication. This simultaneously normalizes all three voltages to compensate for AC line voltage and miscellaneous other variations. Although this “one-size-fits-all” manual voltage regulation technique is functional, it is a coarse technique that compromises instrument transconductance measurement accuracy.

To be fair, most (if not all) tube testers of that era employed this same (or similar) technique. The problem was that with 1960s-vintage vacuum tube technology, precise electronic voltage regulation required elaborate circuitry that was inconsistent with modest instrument pricing. Having a manual adjustment to normalize the various voltages as discussed above was thus an expedient technique that facilitated a reasonable cost-performance trade-off. With the benefit of modern technology, however, the dimensions of this trade-off can be significantly reduced for enhanced instrument precision.

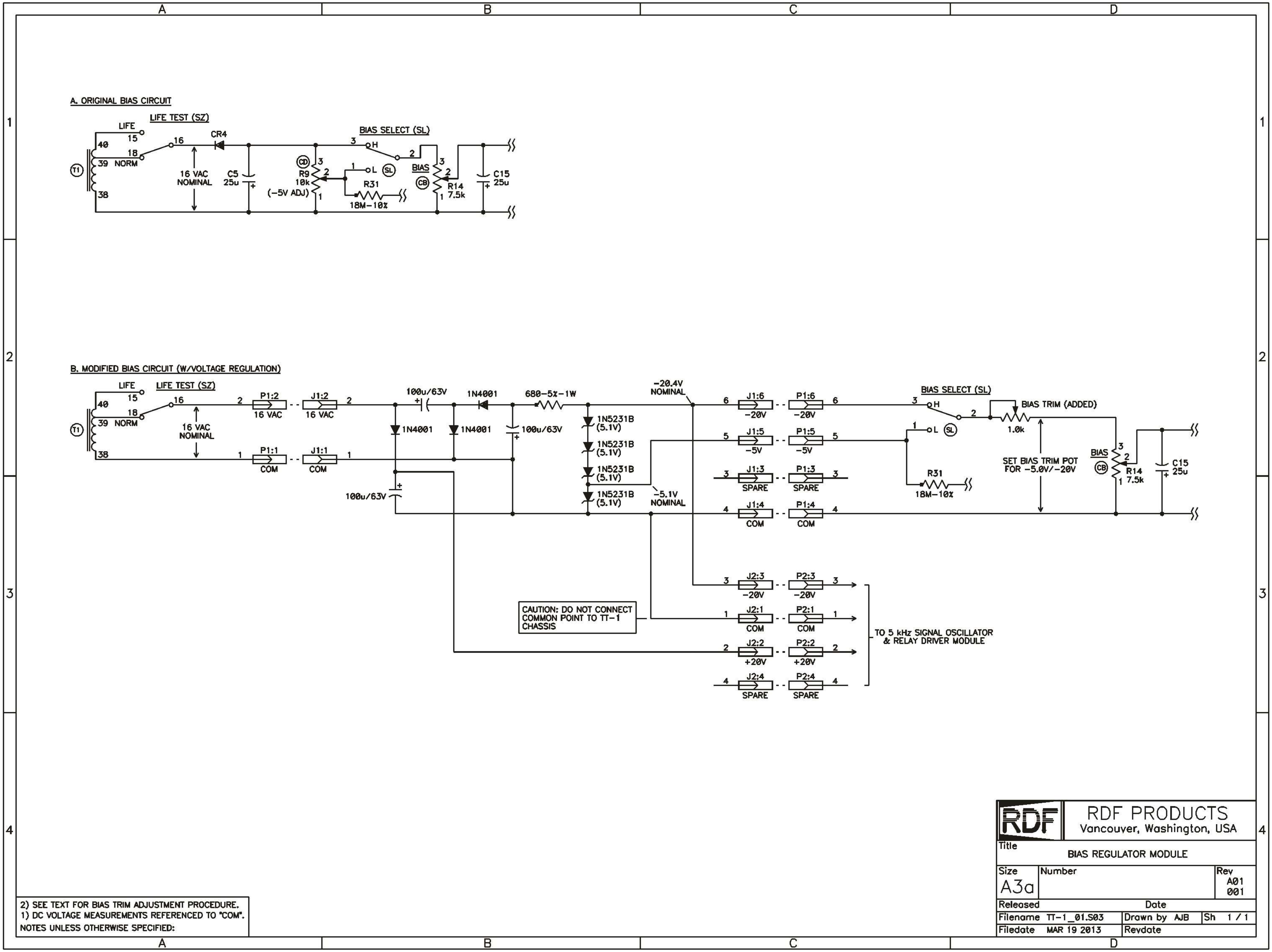
Of the three tube operating voltages, transconductance is most affected by grid bias voltage variations, significantly affected by filament voltage variations, and least affected by plate voltage variations. (For most tubes, modest changes in plate or screen voltages have only modest effect on transconductance.) The theme underlying this Section is that adding precise electronic voltage regulation where possible allows the Set Line control to focus more narrowly (and thus more precisely) on the filament voltage with the result that transconductance measurement accuracy is improved.

More specifically, the addition of a relatively simple grid bias voltage regulator allows the Set Line control to be used exclusively for precisely setting the more critical filament voltage. Although this results in a wider plate voltage variation, this does not significantly affect transconductance measurement accuracy since modest plate voltage variations do not significantly change tube transconductance.

These and related topics are discussed in more detail in the paragraphs that follow.

B. GRID BIAS VOLTAGE REGULATION

The TT-1A grid bias supply circuit is very conventional and illustrated schematically in Figure 1a. As per that schematic, the nominal 16 VAC voltage appearing between T1 pins 38-39 is applied to a half-wave rectifier comprising rectifier diode CR4 and filter capacitor C5. The rectifier output is nominally -20 VDC with the Set Line control properly adjusted (although in both of my TT-1As this voltage was actually somewhat low).


A. ORIGINAL BIAS CIRCUIT
LIFE TEST (SZ)
LIFE
15
16
18
NORM
40
39
38
T1
CR4
16 VAC NOMINAL
C5 25u
CD
R9 10k (-5V ADJ)
BIAS SELECT (SL)
H
L
SL
R31 18M-10%
BIAS
CB
R14 7.5k
C15 25u
B. MODIFIED BIAS CIRCUIT (W/VOLTAGE REGULATION)
P1:2 16 VAC
J1:2 16 VAC
P1:1 COM
J1:1 COM
100u/63V
1N4001
680-5%-1W
1N5231B (5.1V)
-20.4V NOMINAL
-5.1V NOMINAL
J1:6 -20V
P1:6 -20V
J1:5 -5V
P1:5 -5V
J1:3 SPARE
P1:3 SPARE
J1:4 COM
P1:4 COM
BIAS TRIM (ADDED)
1.0k
SET BIAS TRIM POT FOR -5.0V/-20V
J2:3 -20V
P2:3 -20V
J2:1 COM
P2:1 COM
J2:2 +20V
P2:2 +20V
J2:4 SPARE
P2:4 SPARE
TO 5 kHz SIGNAL OSCILLATOR & RELAY DRIVER MODULE
CAUTION: DO NOT CONNECT COMMON POINT TO TT-1 CHASSIS
NOTES UNLESS OTHERWISE SPECIFIED:
1) DC VOLTAGE MEASUREMENTS REFERENCED TO "COM".
2) SEE TEXT FOR BIAS TRIM ADJUSTMENT PROCEDURE.
RDF PRODUCTS
Vancouver, Washington, USA
Title
BIAS REGULATOR MODULE
Size A3a
Number
Rev A01 001
Released
Date
Filename TT-1_01.S03
Drawn by AJB
Sh 1 / 1
Filedate MAR 19 2013
Revdate

F ure 1

R9 is adjusted to produce a lower bias voltage (-5.0 VDC nominal). The front-panel H/L (High/Low) Bias Select switch (SL) is then used to select either the high or low bias output voltage depending upon the requirements of the specific tube under test. Finally, the front-panel Bias potentiometer (R14) is then appropriately set (on a rotational scale of 0-50) for the specified tube grid bias.

R14 has a nominally linear taper (i.e., with the low bias output selected, an R14 setting of 10 nominally corresponds to a -1.0 volt bias output). I did observe, however, that there are noticeable deviations from linearity.

The function of the Life Test switch in this bias circuit is more subtle and requires some explanation. Among other things, the Life Test switch lowers the filament voltage by approximately 10% when moved from its Normal to Life Test position. Note that this Life Test switch has multiple sections and those sections that lower the filament voltage are not illustrated in Figure 1a. Also note that the TT-1A schematic, contrary to what one would expect, illustrates this switch in the Life Test rather than Normal position.

In the absence of compensatory action, setting this switch to Life Test would also lower the bias voltage. To compensate for this undesirable effect, the Life Test switch section illustrated in Figure 1a selects a higher voltage T1 secondary tap (pin 40 rather than 39) when this switch is in the Life Test position. This then restores the AC voltage applied to the bias rectifier to its nominal value (16 VAC) with the result that the bias voltage is nearly unchanged.

With modest effort, the bias circuit can be greatly improved by including a voltage regulator that provides a stabilized bias voltage as illustrated schematically in Figure 1b. My first attempt at this employed an LM2991T low-dropout negative voltage regulator. Unfortunately, this effort soon ran into complications caused by the fact that the nominal 16 VAC transformer output varies widely as a function of both the Set Line control and the actual value of the nominal 115 VAC mains voltage. This was compounded by the fact that the LM2991T has an upper input voltage limit of only 26 VDC, which in turn would have required a pre-regulator for protection.

To avoid all this complexity, I decided instead to employ the simpler (but nonetheless effective) Zener regulator illustrated in Figure 1b. This circuit employs four 1N5231B Zener diodes in series to produce nominal regulated outputs of -5.1 and -20.4 VDC (subsequently trimmed-down to -5.0 and -20.0 volts as discussed below). For best precision, the four Zener diodes should be selected for closely matched knee voltages. If this is not convenient, then the next best move is to select these Zeners from the same batch.

A problem that emerged when designing this regulator was that the original half-wave rectifier did not supply sufficient overhead voltage for the Zeners. To remedy this, the half-wave circuit was discarded in favor of the half-wave voltage doubler illustrated in Figure 1b to produce more output voltage.

The capacitors I used for this voltage doubler were 100 uF/63 VDC units with good ripple current specifications and rated for 105° C operation (Digi-Key P13137-ND/Panasonic EEU-EB1J101). More commonly available 47 uF/50 VDC capacitors, however, are also suitable. Although the voltage doubler and regulator circuit draw more current from T1 than the original half-wave rectifier circuit, this additional current is modest and does not tax T1's current output capabilities.

Testing and adjusting this circuit is straightforward. After carefully checking the wiring for errors, apply power and confirm the -5.1V and -20.4 V nominal Zener voltages with a DVM. Next, pull down the Line Test switch and adjust the Set Line control so that the meter points to the mid-scale Line Check arrow. Once done, connect the DVM between SP2 and SR2 as per the Bias Voltage Supply adjustment procedure presented on page 63 of the TT-1A assembly manual. (This effectively monitors the voltage between pins 1 and 3 of Bias pot R14.) Finally, adjust the added 1.0k Bias Trim pot for a reading of very close to -5.0 VDC when the H/L Bias Select switch is set to low bias and -20.0 VDC when this switch is set to high bias.

If the four Zener diodes are well matched, the same Bias Trim pot setting will be correct (or at least extremely close) for both the low and high bias voltages. If there is a variation, adjust the added Bias Trim pot so as to split the error equally.

This adjustment should be repeated after all the modifications presented in this paper have been successfully completed. When this is done, first set up the TT-1A for a 6AU6 or other typical low-level receiving tube with a 6.3 VAC/300 mA or 12.6 VAC/150 mA filament, plug in the tube, and adjust the Set Line control for the precise nominal filament voltage (discussed in Section IV-D below). This step normalizes the nominal 16 VAC source voltage applied to the bias rectifier so as to provide a more typical DC drive current through the four series Zeners. This in turn mitigates the slight drive voltage sensitivity exhibited by these Zeners. Fortunately, these Zeners are well behaved and exhibit only very small variations as a function of drive current.

This adjustment supercedes the bias voltage adjustments presented on pages 63 and 68 of the TT-1A assembly manual. The addition of this bias regulator makes these tests irrelevant, so they should not be attempted.

Figure 2 - Added Grid Bias Voltage Regulator Module

I built this voltage doubler/regulator module on a 1.7” x 2.7” piece of single-side copper clad perforated prototyping board and mounted it as illustrated in Figure 2. To create more room for this module, I removed R9 (the 10k “CD” pot that had originally been used as the low bias voltage adjustment). *Be certain to heed the safety warning below and not allow any portion of the circuitry to make contact with the TT-1A chassis.*

I mounted the added 1.0k Bias Trim pot directly on the H/L Bias Select switch as illustrated in Figure 3. Note that this pot is staked-down to the switch with Permatex Part 66B RTV to secure it.

**** DANGER ****

> Heathkit designed the TT-1A so that its chassis was electrically isolated from all components to reduce electrical shock hazard. *Do not defeat this important safety feature by making any connections to the chassis when adding the new voltage regulator or when making any other modifications.*

Referring once again to Figure 1b, the reader will notice an additional 1N4007 diode and 100uF/63V capacitor configured as a half-wave rectifier. This circuit produces a +20 VDC unregulated output that is intended for use in the redesigned 5 kHz signal oscillator circuit as discussed in Section V. (The 5 kHz signal oscillator and this grid bias supply have a common ground that allows voltages to be shared.)

The required changes to the TT-1A are fully documented in the Figure 1 schematic.

Figure 3 - Added Bias Trim Pot (mounted on H/L Bias Select switch)

C. PLATE VOLTAGE REGULATION

Although the most elegant solution would include plate voltage regulation, adding suitable regulators would not only be complicated, but would improve transconductance measurement accuracy very modestly. Since this enhancement would be an unnecessary sophistication, this matter is not further addressed in this paper.

D. FILAMENT VOLTAGE REGULATION

Filament voltage regulation is an important voltage stabilization enhancement that can be implemented to improve TT-1A transconductance measurement accuracy. As discussed above, variations in filament voltage produce more change in tube transconductance than comparable percentage variations in plate voltage.

Unfortunately, the Set Line control does not precisely set the filament voltage because the meter Line Check circuit does not directly measure the filament voltage. For best instrument accuracy, the filament voltage should be precisely regulated (either automatically or manually).

The most elegant solution would be to design and build a filament voltage regulator along the lines of the grid bias regulator presented above. Although tube filaments are mostly AC-operated, they can also be DC-operated with no adverse effects. Thus, a DC voltage regulator could, in principle, be built for this purpose. Such a regulator would have to be programmable (i.e., its output voltage would be appropriately selected by the existing TT-1A filament voltage selector switch).

Unfortunately, designing a suitable regulator would be a complicated and demanding task. This regulator would require a selectable output voltage ranging from a low of 0.65 volts to a high of 115 volts. Also, to accommodate all supported tubes, this regulator would need to have an output current capacity of up to 2.5 amperes. In order to handle such a wide voltage range with such a high output current capacity, an elaborate switching regulator would be the most promising choice. While such a regulator could be implemented, the effort required to design and build it would probably discourage most vintage radio enthusiasts (including this one) from attempting such an ambitious undertaking.

Figure 4 - Filament Voltage Voltmeter Probe

A simpler "stripped-down" version of such a regulator could be built with far less effort if only the more commonly-used filament voltages are needed (e.g., 5.0, 6.3, 12.6 volts, etc.). A suitable regulator for these lower voltages could be implemented as a more conventional linear voltage regulator that would be easy to design and build. However, this would diminish the TT-1A's capabilities.

Since I was reluctant to invest the time and effort into building the full-capability filament voltage regulator and also did not want to accept the limitations of the stripped-down alternative, I decided to implement a less elegant (though equally effective) *manual* filament voltage regulation scheme. This scheme relies on the Set Line control to manually set the filament voltage, but employs a precise *external* DVM (rather than the TT-1A panel meter in its Line Check mode) as the filament voltage indicator. Using this technique, the filament voltage can be set precisely to its nominal value. Since DVMs are readily available, this requirement poses no undue burden on most vintage radio enthusiasts.

To make this measurement more convenient, I commandeered the TT-1A AC Line Output receptacle (located on the front-panel upper right corner) for this purpose. Although the intended purpose of this receptacle is to allow convenient connection of a 115 VAC accessory, I anticipated no need for this feature. I therefore disconnected it from the AC line and reconnected it instead to the selected filament voltage output. This selected filament voltage output can be taken from filament voltage selector switch (SM) pin 14 and T1 pin 21. This same voltage is also available at the pilot light test socket (located in the center of the large 7-pin tube socket).

For additional convenience, I constructed a special voltmeter probe for this purpose. As per Figure 4, this probe employs a standard U.S.-style twin-blade AC power plug that conveniently inserts into the TT-1A AC Line Output receptacle. The opposite end of this probe employs two standard banana plugs suitable for connection to most DVMs. *Exercise caution when using this probe - there will be high voltage on the banana plugs in the high voltage filament switch positions.*

On a cautionary note, keep in mind that DVMs are not all created equal. While most DVMs are very precise for DC voltage measurements, they are less accurate for AC. The selected DVM should therefore have good AC specifications. Also, best accuracy is obtained when using the lowest possible DVM scale range suitable for the voltage being measured.

Although a capable DVM should be used, it does not need true-RMS measurement capability (which is a significant cost driver). Since the filament voltage waveform is sinusoidal, less expensive average-reading AC DVMs are fully suitable. (Although these AC DVMs are average-reading, they are calibrated for RMS so long as the waveforms are sinusoidal.

Also, be sure to allow sufficient time for the tube under test to fully warm up when monitoring its filament voltage. Since the filament resistance rises with temperature, the voltage across it will correspondingly rise as well. This will require that the Set Line control be readjusted slightly until the filament reaches its full operating temperature.

On a related note, the manual filament voltage regulation technique using a DVM as proposed above *requires that the TT-1A 5 kHz signal oscillator supply voltages be regulated as discussed in Sections IV-E and V below and that the grid bias supply voltages be regulated as discussed in Section IV-B above.*

To succinctly summarize, the Line Check meter indication (as adjusted by the Set Line control) provides only a proxy indication of the filament voltage that is not very accurate. By monitoring the filament voltage independently and directly with an AC DVM, the Set Line control can be used to precisely (and exclusively) set the filament voltage for improved TT-1A transconductance measurement accuracy.

E. 5 kHz SIGNAL OSCILLATOR VOLTAGE REGULATION

If the steps above are taken (adding a grid bias voltage regulator and manually regulating the filament voltage with the Set Line control using a DVM as a filament voltage monitor as discussed above), the 5 kHz signal oscillator output amplitude will exhibit undesirable variations. This will happen as a result of the fact that the Set Line control varies the oscillator plate and filament voltages. It will therefore be necessary to stabilize the signal oscillator output.

This issue is best solved by replacing the existing 5 kHz vacuum tube signal oscillator with a much cleaner and more modern solid-state implementation as discussed in the following Section. This replacement oscillator employs voltage regulators that make the signal oscillator output impervious to any changes in the Set Line control setting.

Keep in mind that implementing the replacement oscillator is an essential step if the changes described so far are implemented. Readers not desiring to do any one of these improvements should not do any of them at all.

SECTION V - IMPROVING THE 5 kHz SIGNAL OSCILLATOR

A. OVERVIEW

As per my closing comments in VR-004, I noticed that the 5 kHz signal oscillator output was highly non-sinusoidal, meaning that the waveform was harmonic-rich. With such a substantial amount of signal power in these harmonics, the plate choke (L1) impedance presented to the tube under test is different at these harmonic frequencies as compared to the 5 kHz fundamental frequency. Since my two TT-1A signal oscillators exhibited markedly different waveforms (with correspondingly different harmonic spectral outputs), this seemed to be the basis for a reasonable explanation for at least some of the discrepancy in transconductance measurement results between the two instruments.

In addition to signal oscillator waveform distortion, I also noticed that both oscillator frequencies were significantly in error to the high side. Furthermore, there is no oscillator frequency adjustment.

To correct these shortcomings, I replaced the existing 5 kHz signal oscillator with one having the following improvements:

1. A low-distortion sinusoidal output.
2. A stable operating frequency very close to 5 kHz.
3. A stable output amplitude independent of the Set Line control setting.

Designing a stable, sinusoidal audio oscillator is not a trivial task. Although stable 5 kHz signal sources are easy to design and build, achieving a clean sinusoidal output is considerably more difficult.

In the paragraphs that follow, the existing TT-1A signal oscillator circuit is discussed, followed by four alternative approaches. Of these four alternatives, specific circuitry is provided for two of them. Additional topics are also addressed.

B. EXISTING TT-1A SIGNAL OSCILLATOR

The existing TT-1A signal oscillator is an Armstrong type that relies upon inductive (transformer-coupled) feedback via transformer T2. An additional winding (the GRN/YEL wires as per the TT-1A schematic) is used as the signal output. This output is then applied to a precision resistive voltage divider with one of its four output voltages selected by the Signal selector switch (SO). This selected output voltage is then applied to the grid of the tube under test. 10k pot R24 (CG) sets the signal drive level to the voltage divider during instrument calibration.

The oscillator frequency is determined by various circuit capacitances resonating with T2's winding inductance. Although the composition of the resonant LC tank circuit is not clear from the schematic, the distorted (non-sinusoidal) output waveform is a strong indication that the tank Q is very low.

In case the reader is wondering if there is something “magic” about 5 kHz, the answer is no. If I had to make a guess, it would be that the Heathkit engineers chose 5 kHz as a mid-range audio frequency that was easy to work with (i.e., not too high, but far enough removed from the 50/60 Hz power supply ripple frequency for easy ripple voltage rejection filtering).

C. REPLACEMENT ALTERNATIVE #1 - HIGH-Q LC OSCILLATOR

I present this alternative mostly as a matter of interest since of these four alternatives it is the one that most resembles the original TT-1A signal oscillator circuit. I do not present an actual circuit since an LC oscillator is no longer considered to be a modern solution for an audio signal source (primarily due to the bulk and expense of the inductor, coupled with the fact that modern technology offers easier and less expensive solutions).

The biggest advantage of the high-Q LC oscillator is that it inherently provides a very clean sinusoidal output if properly designed. More specifically, the high-Q tank circuit serves as a selective bandpass filter that effectively attenuates harmonics. To take advantage of this, the output should be loosely coupled directly from this tank circuit.

A major design issue would be finding a suitable inductor. Back in the 1960s, high-Q 88 mH telephone toroids were readily available and were widely used by radio amateurs for audio oscillators, CW audio filters, and radio teletype audio FSK demodulators. Unfortunately, these toroids are no longer in production and are likely hard to find.

If I were to pursue this alternative, my approach would be to employ a Colpitts oscillator (to avoid the requirement for a tapped inductor) using a JFET or operational amplifier as the active element. I would then loosely couple the signal from the high-Q tank circuit into a suitable high input impedance buffer amplifier. A suitable tank inductor could likely be found among the large number of inductors available for switching power supplies.

D. REPLACEMENT ALTERNATIVE #2 - STABILIZED WIEN-BRIDGE OSCILLATOR

The stabilized Wien-bridge circuit of Figure 5 (taken from the “IC Op-Amp Cookbook” by Walter J. Jung) is a simple and elegant solution for a replacement 5 kHz signal oscillator. This circuit relies on the temperature sensitive resistance of a tungsten incandescent lamp filament to regulate the feedback and prevent signal clipping. When properly adjusted, this circuit produces a very clean sinusoidal output and is sufficiently temperature-stable if good quality components are used for R1/R2 and C1/C2 (i.e., 1% metal film resistors and Mylar capacitors). R1 and R2 can be trimmed slightly to set the oscillator on frequency. The only downside to this circuit is that lamp L1 is out of production and difficult to find. The only additional information I have on this lamp is that it is an Eldema 1869 that draws 14 mA at 10 volts. I believe that this lamp was a low-current dial or indicator light that was used prior to the advent of more power efficient LEDs. While this circuit might be adaptable to more readily available incandescent lamps, this would require investigation and experimentation.

Aside from the lamp issue, this circuit is attractive based on its simplicity and good performance. Although it is a little tricky to initially adjust, once done it is “set-and-forget”. Note that the circuit as illustrated requires a dual-supply operational amplifier. Single-supply operation is possible with suitable modifications. In either case, regulated supply voltages should be used. This required voltage regulation is easily obtained using inexpensive 3-terminal IC voltage regulators.

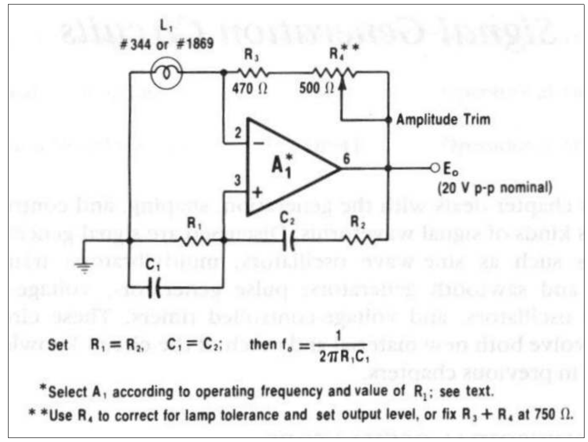


Figure 5 - Wien-Bridge Oscillator (from “IC Op-Amp Cookbook”, 3rd Edition by Walter G. Jung)

E. REPLACEMENT ALTERNATIVE #3 - DIGITALLY-GENERATED SIGNAL SOURCE

The most frequency-stable oscillators are ones that are crystal-controlled. Although crystals are readily available down to 30 kHz or so, 5 kHz crystals are very rare (if they even exist at all). A better solution is to build an oscillator using an inexpensive and readily available microprocessor clock crystal and then digitally divide it down to a frequency near 5 kHz.

The easiest way to construct such a crystal oscillator is to use an HC4060 CMOS 14-stage binary counter/divider and oscillator. This IC contains a convenient logic gate oscillator circuit that is easy to use and suitable for most microprocessor crystals (up to approximately 10 MHz, depending upon the IC manufacturer). The oscillator output is then divided down using cascaded binary counters to provide signal outputs with division ratios of 16, 32, 64, 128, 256, 512, 1024, 4096, 8192, and 16384. The HC4060 data sheet is readily available on-line.

Using this technique, a standard 5.000 MHz microprocessor crystal, for example, yields an output of 4.883 kHz when divided down by 1024. This is an excellent match for the nominal 5 kHz TT-1A signal oscillator design frequency.

The difficulty with this scheme is that the output is a square wave rather than the desired sinusoidal waveform. Fortunately, this can be remedied using an active lowpass filter. Unfortunately, this active filter would have to be somewhat elaborate to produce a clean sinusoidal waveform.

A more elegant approach is to use an end-around shift register configured as a “walking-ring” counter as part of the frequency divider chain. This allows the use of shift register output weighting resistors to produce a much better piecewise approximation of a sinusoidal waveform than the hard square wave output produced by straight binary dividers.

This concept is well-illustrated in Figure 6 (taken from the “CMOS Cookbook” by Don Lancaster and Howard M. Berlin). As per that figure, a four-stage end-around shift register is configured to divide the clock input frequency by a factor of eight. The output is taken through three appropriately selected weighting resistors connected to the Q output of the first three flip-flops. With the weighting resistor values illustrated, the summed output is an optimal piecewise approximation of a sinusoidal waveform. The advantage of this walking-ring configuration is that much less subsequent lowpass filtering is required to produce a clean sinusoidal waveform (i.e., one without the waveform stair-steps) than would be the case for a hard square wave input. In effect, the walking-ring counter is a simple digital-to-analog converter.

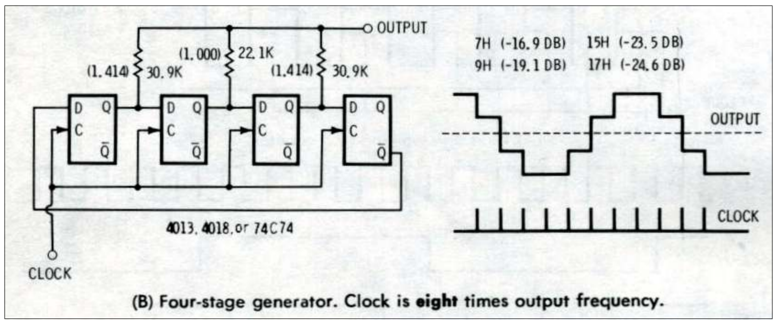
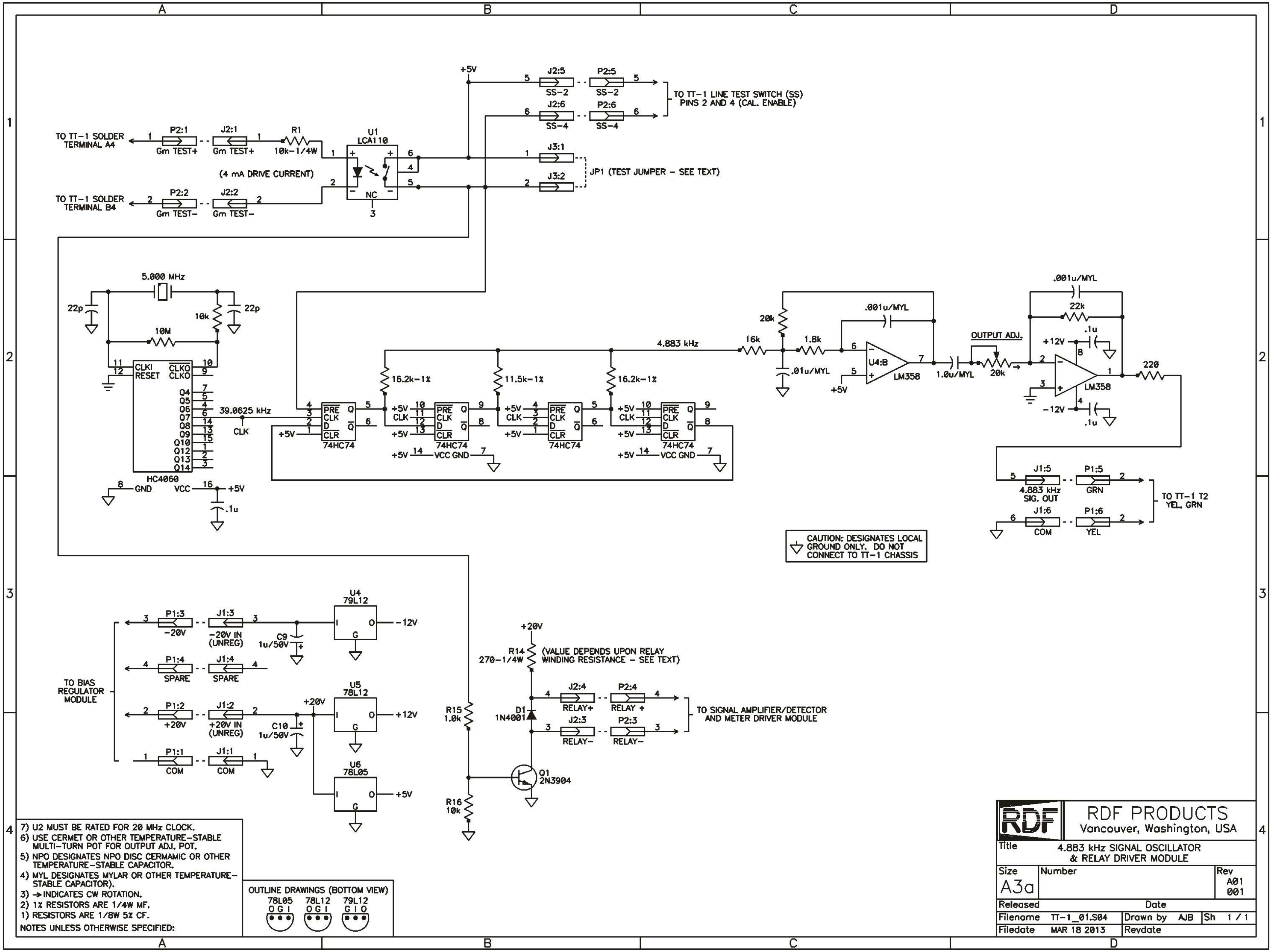


Figure 6 - CMOS Walking-Ring Counter/Divider
(from "CMOS Cookbook, 2nd Edition
by Don Lancaster, revised by Howard M. Berlin)

The complete circuit (including the active lowpass filter) is illustrated in Figure 7. On the positive side, this circuit provides a stable sinusoidal output, requires no adjustments (aside from setting the desired output amplitude), is repeatable, straightforward, and uses readily available and inexpensive parts. The disadvantage is that it has a high parts count.

The parts count can be reduced by substituting a single 4-bit binary counter IC for the two 74HC74s (I used the two 74HC74s as a matter of convenience because they were on-hand). Many years ago I used the 4018 for this purpose, but this IC is no longer in production. The 74HC161appears to be a more modern substitute, although I have no experience with it. By making this substitution and using a dual operational amplifier for the active filter and output driver, the entire circuit can be implemented with three IC packages.

Although I did actually build this circuit, it should be considered as preliminary since I subsequently chose the superior firmware-derived signal source discussed below and thus never actually integrated the walking-ring signal source into my TT-1A. Readers choosing this solution should therefore validate circuit performance carefully and be prepared to make any necessary amendments.


TO TT-1 SOLDER TERMINAL A4
TO TT-1 SOLDER TERMINAL B4
P2:1 Gm TEST+
J2:1 Gm TEST+
P2:2 Gm TEST-
J2:2 Gm TEST-
R1 10k-1/4W
(4 mA DRIVE CURRENT)
U1 LCA110
NC
+5V
J2:5 SS-2
P2:5 SS-2
J2:6 SS-4
P2:6 SS-4
TO TT-1 LINE TEST SWITCH (SS) PINS 2 AND 4 (CAL. ENABLE)
J3:1
J3:2
JP1 (TEST JUMPER - SEE TEXT)
5.000 MHz
22p
10k
10M
CLKI RESET CLKO CLKO
HC4060
GND VCC
39.0625 kHz
CLK
.1u
PRE CLK D CLR Q
74HC74
VCC GND
16.2k-1%
11.5k-1%
4.883 kHz
16k
20k
1.8k
.01u/MYL
.001u/MYL
U4:B
LM358
1.0u/MYL
OUTPUT ADJ.
22k
+12V
-12V
220
J1:5 4.883 kHz SIG. OUT
P1:5 GRN
J1:6 COM
P1:6 YEL
TO TT-1 T2 YEL, GRN
CAUTION: DESIGNATES LOCAL GROUND ONLY. DO NOT CONNECT TO TT-1 CHASSIS
P1:3 -20V
J1:3 -20V IN (UNREG)
C9 1u/50V
U4 79L12
P1:4 SPARE
J1:4 SPARE
TO BIAS REGULATOR MODULE
P1:2 +20V
J1:2 +20V IN (UNREG)
C10 1u/50V
U5 78L12
P1:1 COM
J1:1 COM
U6 78L05
+20V
R14 270-1/4W
(VALUE DEPENDS UPON RELAY WINDING RESISTANCE - SEE TEXT)
J2:4 RELAY+
P2:4 RELAY +
J2:3 RELAY-
P2:3 RELAY-
TO SIGNAL AMPLIFIER/DETECTOR AND METER DRIVER MODULE
D1 1N4001
R15 1.0k
R16 10k
Q1 2N3904
7) U2 MUST BE RATED FOR 20 MHz CLOCK.
6) USE CERMET OR OTHER TEMPERATURE-STABLE MULTI-TURN POT FOR OUTPUT ADJ. POT.
5) NPO DESIGNATES NPO DISC CERMAMIC OR OTHER TEMPERATURE-STABLE CAPACITOR.
4) MYL DESIGNATES MYLAR OR OTHER TEMPERATURE-STABLE CAPACITOR).
3) → INDICATES CW ROTATION.
2) 1% RESISTORS ARE 1/4W MF.
1) RESISTORS ARE 1/8W 5% CF.
NOTES UNLESS OTHERWISE SPECIFIED:
OUTLINE DRAWINGS (BOTTOM VIEW)
78L05 O G I
78L12 O G I
79L12 G I O
RDF PRODUCTS
Vancouver, Washington, USA
Title 4.883 kHz SIGNAL OSCILLATOR & RELAY DRIVER MODULE
Size A3a
Number
Rev A01 001
Released
Date
Filename TT-1_01.S04
Drawn by AJB
Sh 1 / 1
Filedate MAR 18 2013
Revdate

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F. REPLACEMENT ALTERNATIVE #4 - FIRMWARE-DERIVED SIGNAL SOURCE

1. OVERVIEW

An even more modern and elegant approach employs a small microprocessor to generate the desired 5 kHz sinusoidal waveform using firmware. In effect, this microprocessor is programmed as a “smart” digital-to-analog converter. Although this approach is more sophisticated and technically advanced than the walking-ring counter technique presented above, it actually has a smaller parts count and is easier to implement. The most difficult part is writing the code, but I have posted both the source code and the compiled hex file on the N6DC vintage radio website for the benefit of interested readers.

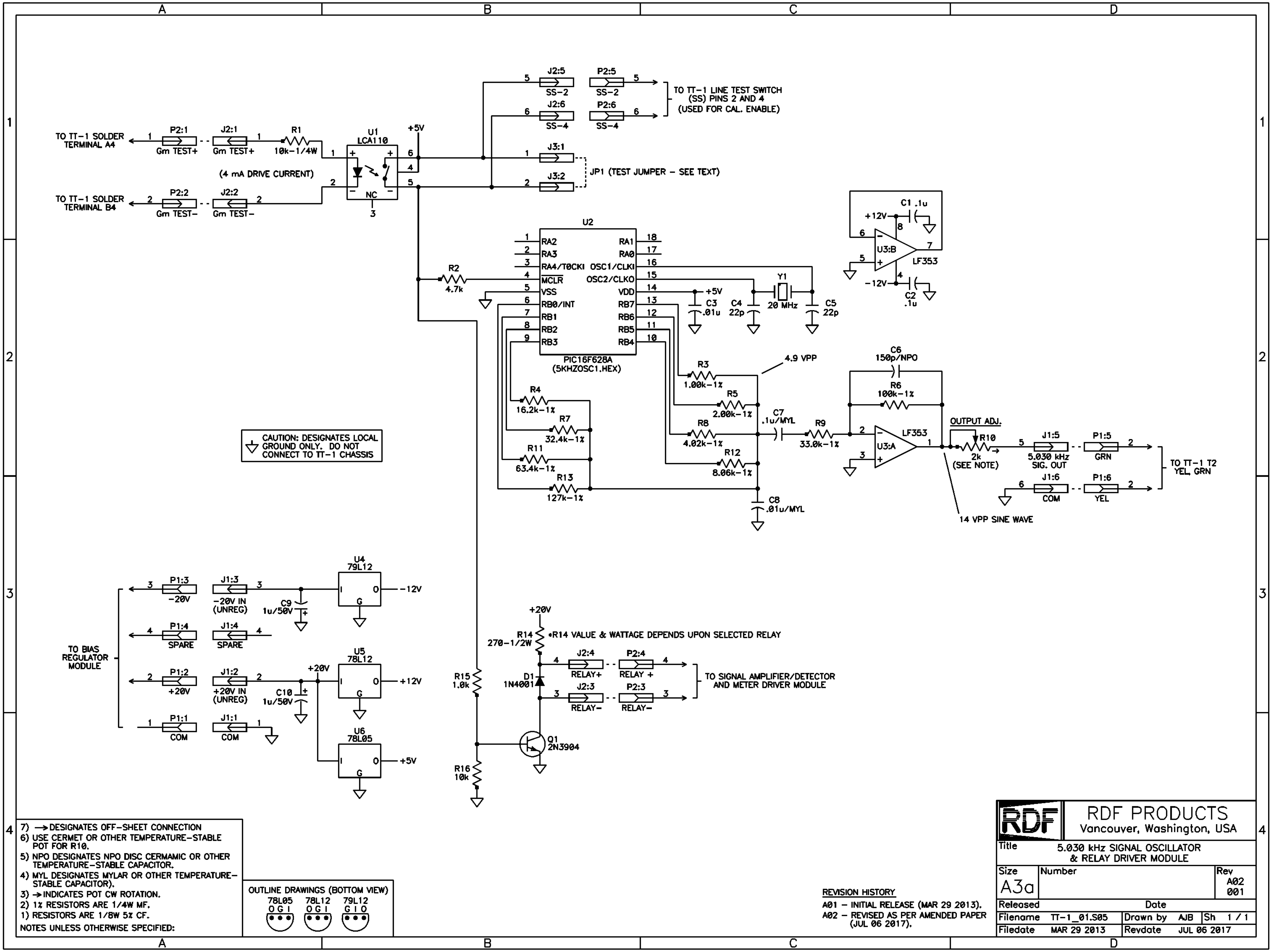
The complete circuit of the new 5 kHz Signal Oscillator and Relay Driver Module is illustrated in Figure 8. The heart of this circuit is an inexpensive and readily available Microchip PIC 16F628A microprocessor (or “PIC” for short). This circuit actually has many functional similarities to the walking-ring counter approach discussed above. First, both circuits employ a stable crystal-controlled oscillator as the primary signal source. Second, both circuits employ digital frequency division to obtain the desired 5 kHz nominal output. Finally, both circuits employ weighting resistors to produce an optimal sinusoidal waveform piecewise approximation.

The major difference is in the implementation of the digital-to-analog converter. In the walking-ring counter technique discussed above, the digital-to-analog converter is implemented using weighting resistors connected to the flip-flop outputs in an end-around shift register. In the firmware approach, this end-around shift register is *emulated* with a software program loop.

For the benefit of readers with limited experience in computer and microprocessor programming, most programs have a “main loop”. The microprocessor “pointer” (the internal mechanism that determines which line of code is to be executed) begins execution at the first line of code in the listing and then moves down this listing one line of code at a time and appropriately executes each program instruction. Although there may be branching and sub-loops, the pointer eventually is redirected back to the beginning of the listing and the process is repeated (thus giving rise to the term “loop”).

Referring to Figure 8, eight of the PIC “ports” (RB0 through RB7) are configured as outputs that can be instructed by the program to output either a logic low (0V) or a logic high (+5V). These ports connect to a network of eight summed weighting resistors in a fashion very similar to that of the walking-ring counter discussed above. The weighting factors are different, however, with the resistor values incremented in a binary sequence (i.e., each resistor in the network very nearly twice as large in value as the preceding resistor).

With eight weighting resistors whose values are incremented in a binary sequence, a total of 256 discrete summed output voltage values are possible (ranging from zero volts when ports RB0-RB7 are all low to maximum when these ports are all high). This is 64 times better than the four discrete summed output voltage values that can be obtained from the walking-ring counter circuit presented above. This yields much smaller output waveform stair-steps that are much easier for the subsequent lowpass filter to smooth out.



RDF PRODUCTS
Vancouver, Washington, USA
Title 5.030 kHz SIGNAL OSCILLATOR & RELAY DRIVER MODULE
Size A3a
Number
Rev A02 001
Released
Date
Filename TT-1_01.S05
Drawn by AJB
Sh 1 / 1
Filedate MAR 29 2013
Revdate JUL 06 2017
TO TT-1 SOLDER TERMINAL A4
TO TT-1 SOLDER TERMINAL B4
P2:1 Gm TEST+
J2:1 Gm TEST+
P2:2 Gm TEST-
J2:2 Gm TEST-
R1 10k-1/4W
(4 mA DRIVE CURRENT)
U1 LCA110
NC
+5V
J2:5 SS-2
P2:5 SS-2
J2:6 SS-4
P2:6 SS-4
TO TT-1 LINE TEST SWITCH (SS) PINS 2 AND 4 (USED FOR CAL. ENABLE)
J3:1
J3:2
JP1 (TEST JUMPER - SEE TEXT)
U2
PIC16F628A (5KHZOSC1.HEX)
RA2 RA3 RA4/T0CKI MCLR VSS RB0/INT RB1 RB2 RB3
RA1 RA0 OSC1/CLKI OSC2/CLKO VDD RB7 RB6 RB5 RB4
R2 4.7k
C3 .01u
C4 22p
Y1 20 MHz
C5 22p
C1 .1u
C2 .1u
U3:B LF353
+12V
-12V
4.9 VPP
R3 1.00k-1%
R5 2.00k-1%
R8 4.02k-1%
R12 8.06k-1%
R4 16.2k-1%
R7 32.4k-1%
R11 63.4k-1%
R13 127k-1%
C7 .1u/MYL
C8 .01u/MYL
R9 33.0k-1%
C6 150p/NPO
R6 100k-1%
U3:A LF353
OUTPUT ADJ.
R10 2k (SEE NOTE)
14 VPP SINE WAVE
J1:5 5.030 kHz SIG. OUT
P1:5 GRN
J1:6 COM
P1:6 YEL
TO TT-1 T2 YEL, GRN
CAUTION: DESIGNATES LOCAL GROUND ONLY. DO NOT CONNECT TO TT-1 CHASSIS
U4 79L12
U5 78L12
U6 78L05
-12V
+12V
+5V
+20V
P1:3 -20V
J1:3 -20V IN (UNREG)
C9 1u/50V
P1:4 SPARE
J1:4 SPARE
P1:2 +20V
J1:2 +20V IN (UNREG)
C10 1u/50V
P1:1 COM
J1:1 COM
TO BIAS REGULATOR MODULE
R14 270-1/2W
*R14 VALUE & WATTAGE DEPENDS UPON SELECTED RELAY
D1 1N4001
J2:4 RELAY+
P2:4 RELAY +
J2:3 RELAY-
P2:3 RELAY-
TO SIGNAL AMPLIFIER/DETECTOR AND METER DRIVER MODULE
Q1 2N3904
R15 1.0k
R16 10k
7) → DESIGNATES OFF-SHEET CONNECTION
6) USE CERMET OR OTHER TEMPERATURE-STABLE POT FOR R10.
5) NPO DESIGNATES NPO DISC CERMAMIC OR OTHER TEMPERATURE-STABLE CAPACITOR.
4) MYL DESIGNATES MYLAR OR OTHER TEMPERATURE-STABLE CAPACITOR).
3) → INDICATES POT CW ROTATION.
2) 1% RESISTORS ARE 1/4W MF.
1) RESISTORS ARE 1/8W 5% CF.
NOTES UNLESS OTHERWISE SPECIFIED:
OUTLINE DRAWINGS (BOTTOM VIEW)
78L05 O G I
78L12 O G I
79L12 G I O
REVISION HISTORY
A01 - INITIAL RELEASE (MAR 29 2013).
A02 - REVISED AS PER AMENDED PAPER (JUL 06 2017).

F ure 8

Furthermore, the waveform update rate is also much faster. Using the 20 MHz PIC clock oscillator as illustrated in Figure 8 for a 5.030 kHz output, the waveform is updated 96 times per waveform cycle. This is 12 times better than the walking-ring counter with an update rate of only eight times per waveform cycle.

With these near-astronomical improvements in step amplitude and temporal resolutions, the end result is an excellent piece-wise approximation of a sinusoidal waveform. Only a simple dual-section RC lowpass filter is necessary to filter out these residual stair-steps (in contrast to the much more elaborate active filter required for the walking-ring counter output). As can be seen from the Figure 9 oscilloscope photo, its output waveform is a very clean sinusoid with just a residual trace of stair-stepping (which is filtered-out in subsequent signal processing circuitry).

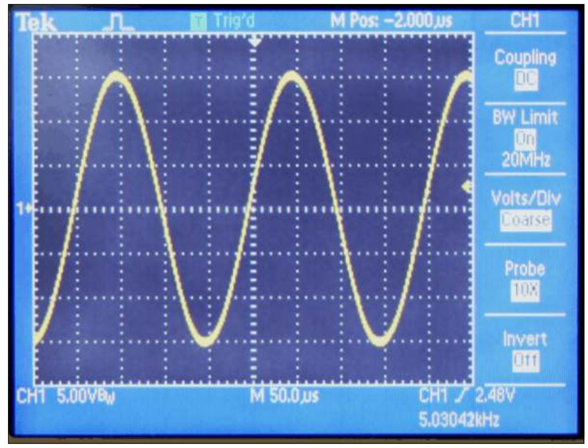


Figure 9 - 5.030 kHz Firmware-Derived Signal Oscillator Output Waveform (measured between T2-RED and T2-BLK)

Since the advantages of this firmware-derived signal source approach are so overwhelming, this is the technique that I chose to implement as the new signal source for my TT-1B. Even so, this may not be the best approach for all readers. In order to implement this firmware-derived approach, readers will need a suitable PIC programmer and must be familiar with its use. To aid interested readers, I have placed the source code and the compiled hex file along with my PIC programmer setups for this firmware on the N6DC vintage radio website. For the benefit of readers not having a suitable PIC programmer, I will send a programmed PIC 16F628A to any domestic USA address in exchange for a stamped self-addressed return mailer and $5.00 cash (to cover costs).

2. DETAILED CIRCUIT DESCRIPTION

Continuing the circuit description in more depth and referring again to Figure 8, the PIC (U2) is the heart of the improved circuit. As per the more general discussion above, it is essentially programmed to function as a digital frequency divider. Fundamentally, it divides down its 20 MHz clock crystal Y1 to 5.030 kHz by means of a software loop as discussed above. By using summed weighting resistors of appropriate values at its eight output ports (RB0-RB7) and with carefully selected sequencing of these ports, an optimal piecewise approximation of a sinusoidal waveform is achieved with minimal stair-stepping.

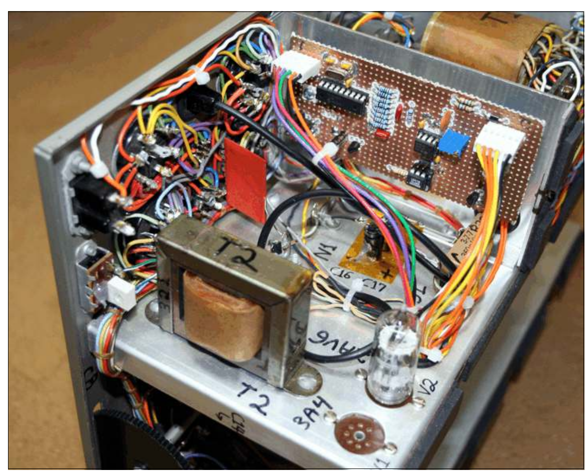


Figure 10 - Added 5.030 kHz Signal Oscillator Module

This stair-step sinusoidal waveform is lowpass filtered at the resistor summing junction with 0.1 uF capacitor C8. The filtered waveform is then applied to operational amplifier U3:A to raise the signal amplitude. 150 pF capacitor C6 works with parallel feedback resistor R6 to provide additional lowpass filtering. Note that both C6, C8, the summed weighting resistors, and the operational amplifier feedback resistors must be temperature-stable to prevent the output amplitude from varying over temperature. To this end, use an NPO ceramic capacitor for C6, a Mylar capacitor for C8, and 1% metal film resistors as indicated. Similarly, a temperature-stable multi-turn cermet type pot should be used at R10 (the output amplitude adjustment).

The circuit is activated by the existing TT-1A Gm Test switch in the same fashion as for the original signal oscillator (i.e., this switch applied plate voltage to the 3A4 oscillator tube, which must be removed to accommodate the improved circuit). Since the original TT-1A oscillator circuit and this module do not share a common ground, opto-isolator switch U1 is employed to convert this switched plate voltage to a level suitable for activating the PIC signal source. With a 10k value of R1, the current drain from the TT-1A plate supply is approximately 4 mA, which matches the plate current of the replaced 3A4 oscillator tube. A relay could be substituted for the opto-isolator switch, but it would have to be a high-sensitivity type with a winding activation current of no more than 5 mA or so.

Since one of the new features of the TT-1B is a front-panel activated calibration mode, the PIC signal source must be capable of being activated in this new calibration mode as well. The TT-1A Line Test switch is commandeered for this purpose (since the Line Test function is no longer needed in the TT-1B) and is connected to this module through J2:5/J2:6 so that it is in parallel with the opto-isolator switch. When the Line Test switch is pulled down, it thus directly activates the PIC signal source. (See Appendix B for a fuller explanation of this new calibration mode.)

Jumper JP1 is a test jumper that plugs into J3:1/J3:2 to allow the PIC 16F628A signal source to be activated locally for test purposes. It directly parallels the Line Test switch (now used as the Calibration mode switch). Although the PIC can also be activated with either the Gm Test or Line Test/Cal switches, these switches are momentary and thus inconvenient for testing and troubleshooting since they must be held down.

For reasons discussed in depth in Section VI, two relays are required in the added Signal Amplifier/Detector and Meter Driver Module that must be activated concurrently with the PIC signal source. This is accomplished with relay driver Q1 and its associated circuitry. Q1 is driven into conduction (tripping the relays) when +5 VDC is applied to the top end of voltage divider R15/R16. This happens whenever the PIC signal source is activated (either by the Gm Test switch, the Line Test/Cal switch, or the JP1 test jumper as discussed above). These two relays (whose windings are parallel-wired) are connected via J2:3/J2:4.

The value of Q1 collector resistor R14 depends upon the relay winding current (both relays in the Signal Amplifier/Detector and Meter Driver Module should be the same, or at least require the same winding current). R14's exact value will depend upon the actual relays that are used, and should be selected to provide the correct relay winding current. It is best that low-current relays be used. Otherwise, it may be necessary to substitute a higher-current transistor (e.g., the metal can 2N2219A) in place of the plastic case 2N3904.

Diode D1 protects Q1 from the inductive kickback voltage spike generated by the relay windings. Although a 1N4001 is illustrated, this diode can be any of the 1N4000 series. The minimum specification for this diode is that it be capable of handling up to 1 ampere of forward current and have a PIV (peak inverse voltage) rating of 100 volts.

The voltages necessary to power the new 5 kHz Signal Oscillator and Relay Driver Module are obtained from the new Bias Regulator Module (see Figure 1), which are connected via J1. Since these voltages are unregulated, standard 3-terminal linear regulators are employed at U4, U5, and U6 to provide regulated voltages of -12 VDC, +5 VDC, and +12 VDC, respectively. Since the current requirements are very modest, these regulators can be the plastic TO-92 low-power versions. Note: The 78/79-series 3-terminal regulators illustrated in Figure 8 require a minimum input overhead voltage of 3.0 volts for conservative design practice to guarantee that they do not fall out of regulation.

To complete this discussion, be sure that the selected PIC 16F628A is rated for a 20 MHz clock. (Some versions are rated only up to 4 MHz.) Addressing packaging, I selected the DIP-18 version and mounted it in a matching DIP-18 socket for convenience of programming and installation.

3. TEST AND ADJUSTMENT

Since the circuit employs precision components, very little adjustment is necessary. Once the circuit wiring has been debugged and verified, first confirm the presence of the +5, +12, and -12 VDC regulated voltages. Next, probe U2-15 with an oscilloscope to confirm the presence of the 20 MHz PIC clock signal. This signal should have an amplitude of at least 3.5 VPP (although it may appear to be less if the oscilloscope bandwidth is under 20 MHz).

Move the oscilloscope probe to the 4.9 VPP test point (at the PIC output resistor summing junction) and confirm that no signal is visible. Still monitoring this test point, pull down the Gm Test switch and confirm that a 5.030 kHz sinusoidal waveform with an amplitude of close to 4.9 VPP appears. Release the Gm Test switch and confirm that the 5.030 kHz sinusoidal waveform disappears. Once confirmed, pull down the Line Test/Cal switch and confirm that the waveform reappears.

Set the Output Adj. pot (R10) to mid-range, move the oscilloscope probe to the 14 VPP test point at U3-1, and then confirm the presence of a sinusoidal waveform with an amplitude of close to 14 VPP. R10 will be readjusted to its final setting in a subsequent procedure.

Finally, check the operation of the relay driver by temporarily bridging D1 with an 820 ohm 1/4 watt resistor (to simulate the relay coil assuming the relay is not yet connected) and probing the D4 anode (i.e., the Q3 collector). With neither the Gm Test nor Line Test switches pulled down, the indicated voltage should be near +20 VDC (i.e, Vcc). With either of these switches pulled down, the indicated voltage should fall to under 0.5 VDC.

4. REQUIRED CHANGES TO THE TT-1A

As per the Figure 8 schematic, the new 5.030 kHz signal oscillator module must be connected to the new grid bias voltage regulator and signal processor modules. However, it must also be wired into the TT-1A so that it can be activated using existing controls. This wiring is discussed below and also illustrated in the marked-up TT-1A schematic of Figure 11 for further clarity. Referring to Figures 8 and 11, these changes are summarized as follows:

a. P2:1/P2:2 (Gm Test+/Gm Test-) must be wired to TT-1A solder terminals A4 and B4, respectively. As per the TT-1A schematic, the Gm Test switch includes a section (pins 3, 5, and 6) that applies plate voltage to the 3A4 signal oscillator tube. This modification uses this same switch section to instead enable opto-isolator switch U1 in the new signal oscillator module. To remedy a subtle timing issue associated with this change, TT-1A C7 must be changed from 20 uF/350 V to 3.9 uF. This replacement capacitor must be rated for least 150 VDC. The 3A4 oscillator tube (V1) must be removed.

b. Similarly, P2:5/P2:6 (SS-2/SS-4) must be wired to TT-1A Line Test switch terminals SS-2 and SS-4, respectively. (SS is actually used as a calibration enable switch in the modified design.) Note the indicated six wire cuts and two added wire jumpers associated with this change.

c. Signal oscillator transformer T2 is still used in the modified design, but in a different fashion. The GRN/YEL winding now serves as the primary winding driven by the new signal oscillator module 5.030 kHz output at P1:5/P1:6 (GRN/YEL). The signal voltage is stepped-up in the T2 grid (secondary) winding (RED/BLK) and applied to the Signal switch and its associated voltage divider resistors. This requires various cuts and jumps as illustrated. Note the cut between the T2 secondary red (lower) terminal and solder terminal R1, which is hard to see in Figure 11. R24, R27, and V3 must be removed.

d. Note that the T2 secondary winding (RED/BLK) is disconnected from all associated components and then wired to the Signal switch, with the RED wire connected to SO-8 and the BLK wire to SO-10. The T2 BRN/BLU winding is disconnected from all associated components.

e. The original 4.7k resistor R33 is replaced by a 33.0k-1% resistor.

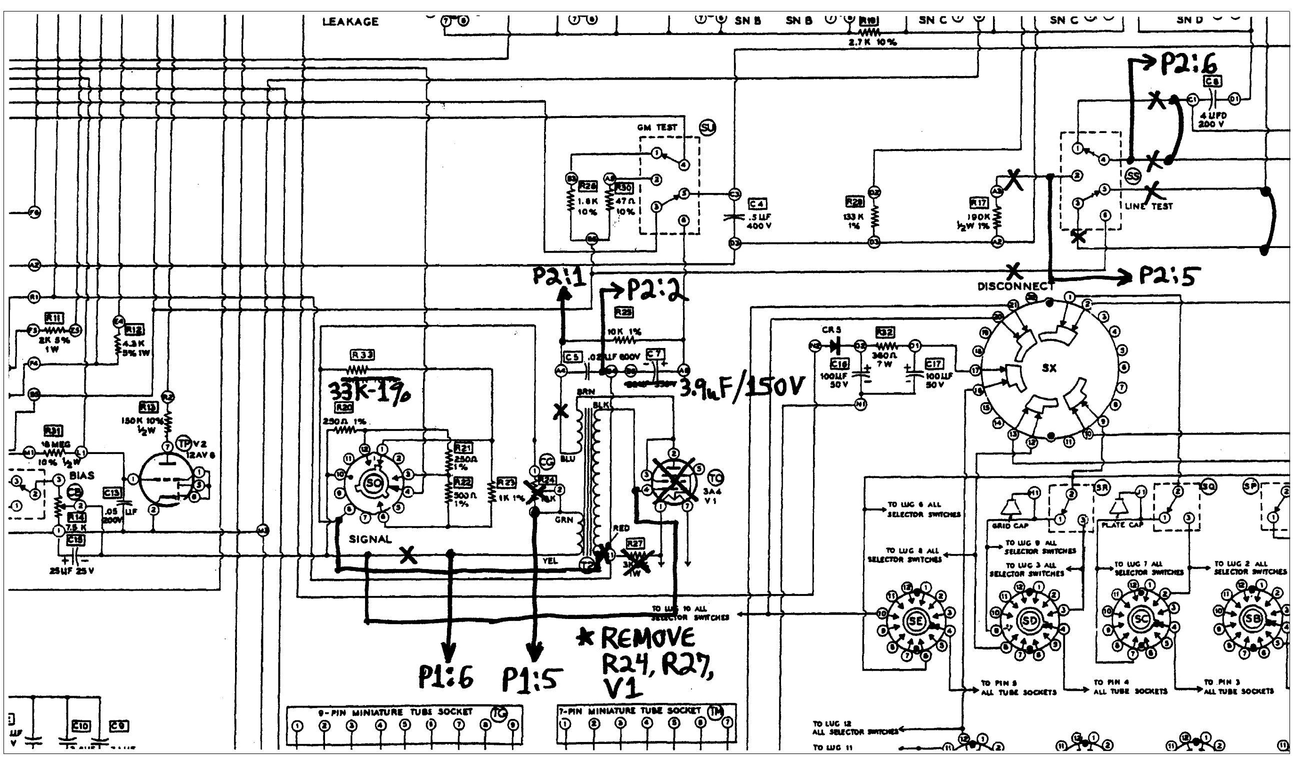


Figure 11 - Wiring the 5.030 kHz Signal Oscillator Module into the TT-1A

SECTION VI - IMPROVING THE SIGNAL PROCESSING AND METERING CIRCUITRY

A. REVIEWING TRANSCONDUCTANCE MEASUREMENT BASICS

Transconductance (or G_m; also referred to as mutual conductance) is formally defined as follows:

$$G_m = \Delta I_p / \Delta E_g \qquad (1)$$

where:

G_m is the transconductance in mhos,
ΔE_g is a small change in tube grid voltage in volts,
ΔI_p is the tube plate current change in amperes resulting from ΔE_g,

and the following conditions apply:

ΔE_g is sufficiently small so that the tube remains within its linear operating region.
The tube plate voltage is held constant.

This latter condition is important. In order for the plate voltage to remain constant, the plate load resistance R_L (or impedance for an AC signal) *implicitly must be zero*. This is more than just an academic subtlety as will be explained shortly.

The ideal tube transconductance measurement functional test setup is illustrated in Figure 12. With the appropriate plate and grid bias voltages applied, a small signal of known amplitude (ΔE_g from above) is applied to the control grid. The resulting AC plate current (ΔI_p from above) is then measured by the AC milliammeter. G_m is then computed as per equation (1).

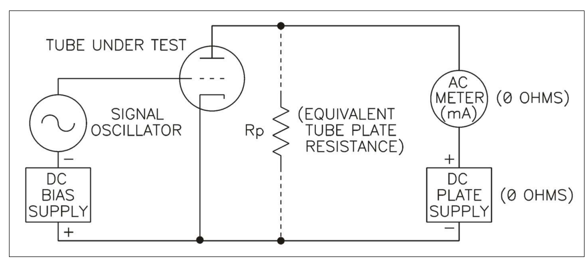
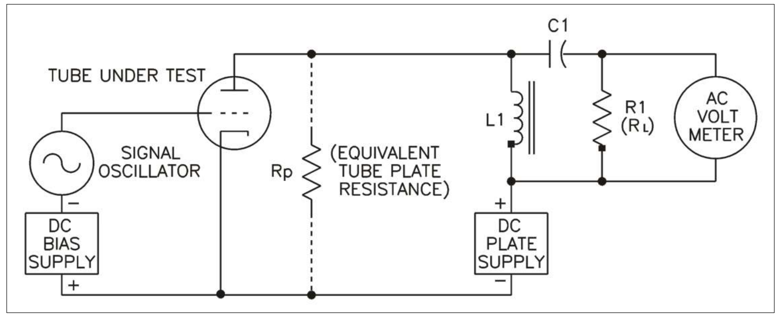


Figure 12 - Ideal Tube Transconductance Measurement Functional Test Setup

In this ideal circuit, both the AC milliammeter and DC plate voltage supply must have no internal resistance. If this condition is not met, the AC plate current flowing through the internal resistance of these components will develop a corresponding AC output voltage swing at the tube plate. This then violates the above condition that the plate voltage must remain constant for equation (1) to be completely accurate (as opposed to being just a useful approximation).

Unfortunately, having a plate load resistance (or impedance) of zero is clearly impractical in any useful circuit since this would not allow any signal to be extracted from the output. A practical tube test circuit must therefore have a *non-zero* value of plate load resistance (R_L) to extract a signal from the output to allow a measurement. This and related concepts are conveniently facilitated by referring to the more practical tube transconductance measurement test circuit of Figure 13. This circuit is based closely on the simplified transconductance test circuit presented in the Heathkit TT-1A operational manual and is typical of most, if not all, transconductance tube testers.

As per Figure 13, an input signal is applied to the tube control grid. This results in the tube plate current being proportionally varied which in turn causes a measurable voltage to be developed across plate load resistor R1.



<u>Figure 13</u> - Practical Tube Transconductance Measurement Functional Test Setup

In reality, the effective load resistance R_L is the plate choke (L1) impedance in parallel with R1 and the AC voltmeter resistance. In a well-designed circuit, the plate choke impedance and AC voltmeter resistances would both be very high compared to R1, or at least well-defined so that we can calculate the composite load impedance. C1's value is such that its capacitive reactance Xc is negligible at the signal oscillator test frequency (5 kHz for the TT-1A), but large at 50/60 Hz so that it helps suppress power supply ripple.

Another useful equation that relates transconductance to voltage gain and load resistance is as follows:

$$A_V = G_m \times R_L \qquad (2)$$

where:

A_V is the voltage gain.
G_m is the transconductance in mhos.
R_L is the plate load resistance in ohms.

and the following conditions apply:

ΔE_g is sufficiently small so that the tube remains within its linear operating region.
R_L is small compared to the tube plate resistance R_p.

If we rearrange equation (2) to solve for G_m, we obtain the following:

$$G_m = A_V/R_L \qquad (3)$$

Referring again to Figure 13, we can compute the tube voltage gain A_V easily enough by measuring the input and output signal voltages (Vi and Vo, respectively) with an AC voltmeter and then simply dividing Vo by Vi. If we know the value of the load resistance R_L, we can then compute the transconductance G_m using equation (3). Figure 13 is the basic circuit used by all tube testers for transconductance measurement.

In reality, equations (2) and (3) are only approximations. To explain, all tubes have a plate resistance R_p (not to be confused with the load resistance R_L) that effectively shunts the load resistance and thus effectively diverts some of the output current. (This equivalent R_p is illustrated in Figure 13 using dotted lines.) In other words, the *effective load resistance is actually R_L in parallel with R_p.*

Equations (2) and (3) are therefore accurate only for cases where R_L is sufficiently small compared to R_p so that the shunting effects of R_p can be ignored for practical measurement purposes. *This is the reason for the condition that R_L be negligible (i.e., that the plate voltage remain constant) for the G_m equation (1).*

This is not strictly an academic issue since the value of R_p can vary over a wide range depending upon the tube type. With sharp cutoff pentodes, R_p is typically in the tens of thousands of ohms and can be safely ignored for test circuits where R_L is typically in the order hundreds of ohms. On the other hand, triodes have much lower plate resistances (in some cases even as low as 1,000 ohms). If the triode plate resistance is not very high compared to R_L (as is often the case in tube tester measurement circuits), significant measurement errors can be expected.

To summarize this discussion, practical tube transconductance measurements can be made using the circuit of Figure 13. With a known value of load resistance R_L, we can compute transconductance by measuring the signal input and output voltages, computing the voltage gain A_V, and then applying equation (3). For good accuracy, the signal input must be sufficiently small so that the tube remains within its linear operating region, and the load resistance R_L must be very small compared to the tube plate resistance R_p.

B. A CRITIQUE OF THE TT-1A Gm MEASUREMENT CIRCUITRY

The TT-1A transconductance measurement circuitry imprecision is predominantly due to the insensitivity and non-linearity of its AC meter. To explain, this AC meter is actually a DC micro-ammeter embedded in a bridge signal rectifier circuit as illustrated schematically in Figure 14. This meter has four terminals, with terminals 1 and 2 being the “AC meter” and terminals 3 and 4 being the “DC meter”. Of course, there is really only one physical meter (a DC meter with an 890 microampere movement and 125 ohm coil resistance).

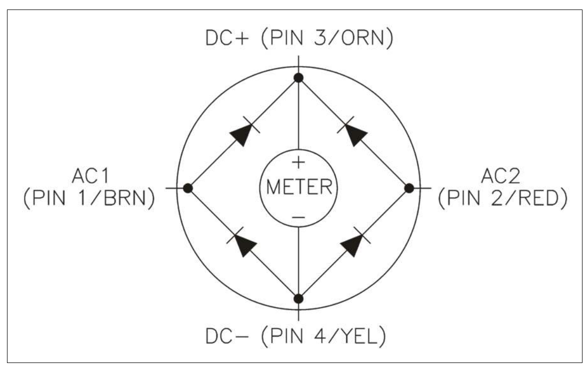


Figure 14 - TT-1A Meter Circuit

Since the rectifier diodes require a certain minimum voltage to reach their conduction threshold, a relatively large AC signal drive voltage is required to obtain accurate measurements. This diode threshold issue is compounded in the TT-1A AC meter since the bridge rectifier circuit requires that the signal must pass through *two* diodes (rather than just one as would be the case for a more conventional half-wave rectifier).

In order to obtain this large AC meter drive voltage, the Heathkit engineers likely had the following two options:

1. Apply a relatively large input drive voltage to the tube grid to obtain a larger tube output voltage. Unfortunately, this violates the condition presented above that this input drive voltage be sufficiently small to avoid output waveform distortion.

2. Employ a relatively large load resistance R_L so as to develop a correspondingly large output voltage swing. Unfortunately, this violates the condition presented earlier in the discussion that R_L be small compared to the tube plate resistance R_p for good accuracy.

It appears that the Heathkit engineers did some of both. Although the circuit output resistance R_L is difficult to determine precisely (it varies significantly with the settings of R1 and R18), I found it to be near 900 ohms when set up for a 6AU6.

Even using these larger than optimum drive voltages and load resistances, I found the metering circuit to be quite non-linear (due to the diodes). This was especially the case in the lower half of the meter scale.

At the end of the day, I believe that the Heathkit engineers had to rely on the Meter control setting (R18) to compensate for all these imprecisions. Although this was likely their best available move to deal with this problem, I believe that this was a somewhat coarse solution that is consistent with Kent Nickerson's belief that the TT-1A transconductance measurement accuracy is only 30%.

A further problem is that given the uncertainties associated with the true value of R_L (comprising L1 in shunt with R1 and the meter circuitry), there is no practical way to actually verify tube transconductance based on objective signal voltage measurements.

Lest it appear that I am judging the Heathkit engineers too critically, I should in all fairness point out that they were no doubt running this project under tight budgetary constraints. Although components and circuitry did exist in 1959 that were capable of more precise signal detection and measurement, this would have required expensive post-amplifiers and more circuitry which would have put the TT-1A out-of-budget for a price conscious market.

If I had to make a guess, it would be that the TT-1A had to employ passive components only for its transconductance measurement function (with the exception, of course, of the 5 kHz signal oscillator) to meet its budgetary constraints. Given this constraint, I think that the Heathkit engineers acquitted themselves well on this project.

C. DESIGN APPROACH FOR ENHANCED TT-1A Gm MEASUREMENT ACCURACY

With the passage of over 50 years, there are much better components and circuits available today (in 2013) than there were in 1959. All of the TT-1A design shortcomings discussed above can be overcome at modest cost using modern solid state design techniques with the new design approach as follows:

1. AC Meter Insensitivity and Non-Linearity - As per the discussion above, this was the primary TT-1A design obstacle that resulted in the various design compromises that degraded accuracy. Addressing first the matter of insensitivity, inexpensive modern operational amplifiers can be used to amplify the tube output signal to any practical level. Addressing the matter of AC meter rectifier non-linearity, precision rectifiers (sometimes referred to in the literature as "super diodes") can be constructed where the rectifier diode is placed in a high-gain operational amplifier feedback loop to wash out the diode non-

linearity. This relatively simple circuit yields nearly ideal rectifier performance.

2. Tube Input Signal Levels - As per the discussion above, the Heathkit engineers had to use excessively high signal drive levels to the tube under test (0.25/0.5/1.0/2.0 VRMS) in order to develop a sufficiently large tube output voltage swing to drive the insensitive meter circuitry. As discussed above, adding a high-gain operational post- amplifier to the tube output circuitry allows practically any desired level of amplification. This in turn allows much smaller tube input signal levels (nominally 0.2/0.4/0.8/1.6 VPP in the new circuit; less than 30% of the original) to ensure that the tube remains within its linear operating region for more accurate transconductance measurement. (With the original circuitry and higher input signal levels, I noticed some distortion on the plate output of a 6AU6 tube under test. Distortion causes gain compression and understates transconductance as a result.)

3. Lower Value of R_L - As per the discussion above, the Heathkit engineers had to use a higher than desired value of R_L in the tube output circuit to develop sufficient voltage swing to drive the insensitive meter circuitry. Once again, with the virtually unlimited gain available in the high-gain operational post-amplifier, we can reduce R_L to a small fraction of its original value with no adverse effects.

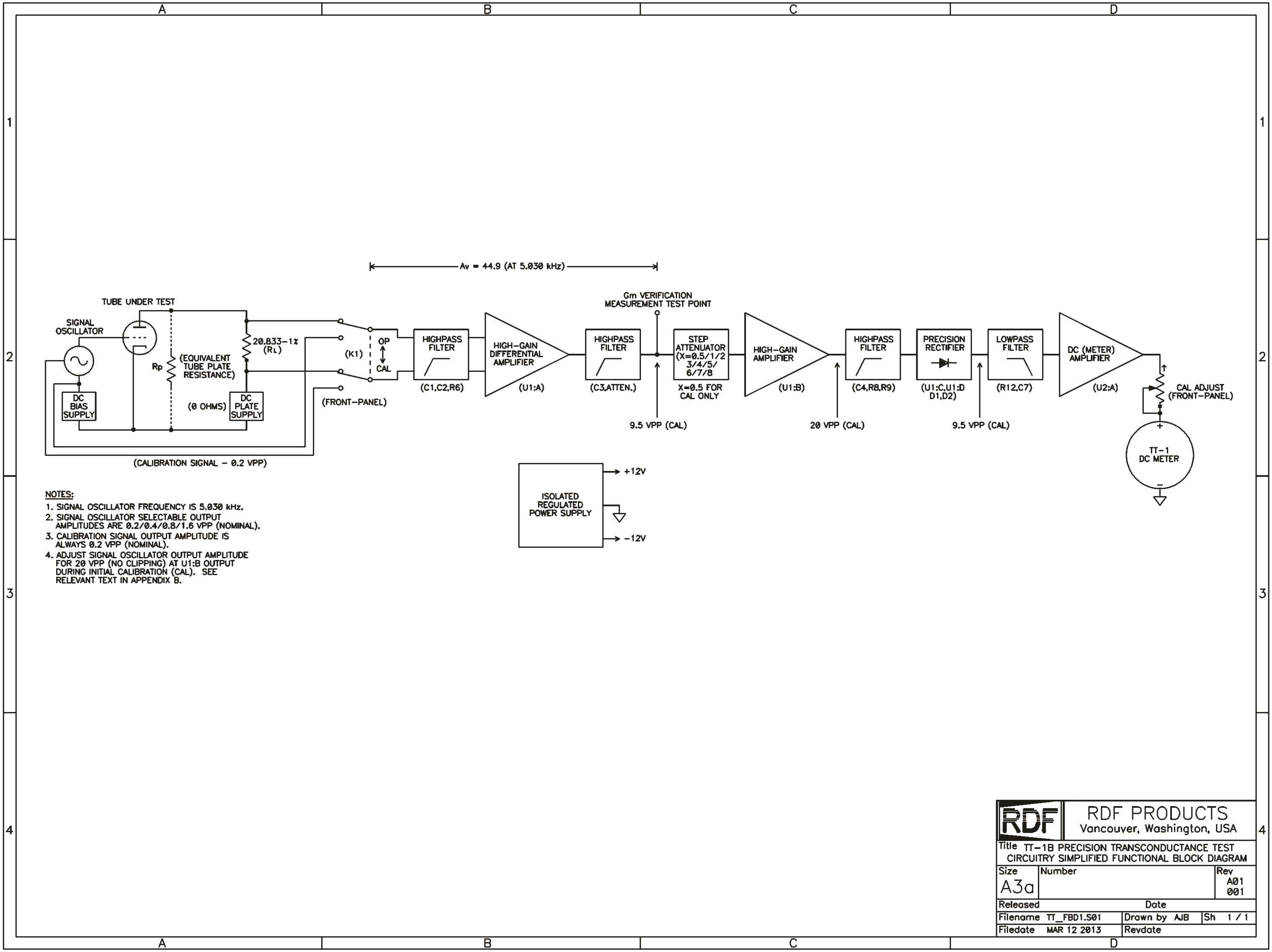
4. Defined Value of R_L - As per the discussion above, the inability to define R_L in the original design makes independent and objective verification of tube transconductance impractical (i.e., TT-1A Gm measurements are essentially "faith-based"). With the new design, this problem is eliminated. Also, transconductance calibration accuracy can be verified on-the-fly with a convenient front-panel test and adjustment procedure.

This implementation of this design approach is presented in detail below.

D. ENHANCED TRANSCONDUCTANCE TEST CIRCUIT FUNCTIONAL DESCRIPTION

With the above design approach in mind and referring to the simplified functional block diagram of Figure 15, the fundamental tube test circuit is similar to the original as illustrated in Figure 13. The difference is that the original plate choke L1 has been discarded and replaced with a 20.833 ohm 1% precision resistor. Although the reason for this unusual value will be explained shortly, the most important thing to understand for the moment is that this new value of R_L is far lower than the combined load impedance of the original plate choke L1 and metering circuitry as illustrated in Figure 13. (This impedance is variable, but is approximately 900 ohms for a 6AU6 tube test setup.)

This change is significant for two important reasons. First, this new value of R_L is constant and unlike in the original circuit does not change with the setting of the TT-1A Meter control (which is disconnected for transconductance measurements in the new design). The tube under test thus always has a constant and well-defined plate load resistance R_L that in turn allows objective and verifiable transconductance measurements based on the tube voltage gain A_V. (The reader will recall from the discussion above that the non-constant and ill-defined plate load impedance of the original circuitry made objective transconductance measurement impractical and unverifiable.)


Av = 44.9 (AT 5.030 kHz)
TUBE UNDER TEST
SIGNAL OSCILLATOR
Rp
(EQUIVALENT TUBE PLATE RESISTANCE)
(0 OHMS)
DC BIAS SUPPLY
DC PLATE SUPPLY
20.833-1% (RL)
(K1)
OP
CAL
(FRONT-PANEL)
(CALIBRATION SIGNAL - 0.2 VPP)
HIGHPASS FILTER
(C1,C2,R6)
HIGH-GAIN DIFFERENTIAL AMPLIFIER
(U1:A)
HIGHPASS FILTER
(C3,ATTEN.)
Gm VERIFICATION MEASUREMENT TEST POINT
9.5 VPP (CAL)
STEP ATTENUATOR (X=0.5/1/2 3/4/5/ 6/7/8
X=0.5 FOR CAL ONLY
HIGH-GAIN AMPLIFIER
(U1:B)
20 VPP (CAL)
HIGHPASS FILTER
(C4,R8,R9)
PRECISION RECTIFIER
(U1:C,U1:D D1,D2)
9.5 VPP (CAL)
LOWPASS FILTER
(R12,C7)
DC (METER) AMPLIFIER
(U2:A)
CAL ADJUST (FRONT-PANEL)
TT-1 DC METER
ISOLATED REGULATED POWER SUPPLY
+12V
-12V
NOTES:
1. SIGNAL OSCILLATOR FREQUENCY IS 5.030 kHz.
2. SIGNAL OSCILLATOR SELECTABLE OUTPUT AMPLITUDES ARE 0.2/0.4/0.8/1.6 VPP (NOMINAL).
3. CALIBRATION SIGNAL OUTPUT AMPLITUDE IS ALWAYS 0.2 VPP (NOMINAL).
4. ADJUST SIGNAL OSCILLATOR OUTPUT AMPLITUDE FOR 20 VPP (NO CLIPPING) AT U1:B OUTPUT DURING INITIAL CALIBRATION (CAL). SEE RELEVANT TEXT IN APPENDIX B.
RDF PRODUCTS
Vancouver, Washington, USA
Title TT-1B PRECISION TRANSCONDUCTANCE TEST CIRCUITRY SIMPLIFIED FUNCTIONAL BLOCK DIAGRAM
Size A3a
Number
Rev A01 001
Released
Date
Filename TT_FBD1.S01
Drawn by AJB
Sh 1 / 1
Filedate MAR 12 2013
Revdate

________F ure 15

Second, this impedance is much lower than the plate resistance of any tube (represented as R_p in Figures 12 and 13). The reader will recall from the previous discussion that when computing transconductance based on voltage gain measurements as per equations (2) and (3), it is important that R_L be much lower than R_p for these equations (which are approximations) to be accurate. With R_L set at only 20.833 ohms, this condition is very well met, even for triodes with very low values of R_p.

A further benefit of using a 20.833 ohm plate load resistor is that this value is sufficiently low that the plate choke L1 used in the original circuit can be eliminated. (This low value resistor does not cause a significant DC voltage drop, thus making the plate choke unnecessary.) This in turn eliminates the uncertainty associated with L1's shunt impedance.

Using a 20.833 ohm resistor for R_L is thus a giant step forward toward improved measurement accuracy. The drawback, however, is that very little signal voltage is developed across this low-value plate load resistor.

Fortunately, this drawback is easily remedied using the high-gain differential post-amplifier as per Figure 15. It is fair to say that the most effective "magic ingredient" that we have at our disposal today that did not exist in 1959 is the modern, inexpensive, IC operational amplifier. Operational amplifiers allow very high gain along with high input and low output impedances, and can accept either single-ended or differential inputs. As per Figure 15, the differentially-driven amplifier receives the low signal voltage developed across the 20.833 ohm plate load resistor R_L (after first passing through a highpass filter to help suppress 50/60 Hz power supply ripple voltage) and amplifies it to a much higher level. Since the differential amplifier has a very high input impedance, it has negligible loading effect on R_L.

Again referring to Figure 15, the differential amplifier output signal is then passed through a second highpass filter for additional 50/60 Hz ripple filtering. The filtered output is then fed to a step attenuator that's function will be explained in greater depth subsequently. The total circuit voltage gain (Av) at this point (the step attenuator input) is close to 45.

The test signal is then passed through another high-gain operational amplifier to further build up the signal amplitude and then passed through a third highpass filter to further suppress residual power supply ripple voltage.

To explain why so many highpass filters are required, the extremely low R_L value (20.833 ohms) results in a very small 5.030 kHz signal output. As a result, the 50/60 Hz plate supply ripple voltage component is significant and must be thoroughly filtered out for accurate measurements. To this end, the circuit includes three simple single-section RC highpass filters (each having 3-dB cutoff frequencies in the 500-600 Hz range) to suppress the 50/60 Hz ripple voltage to a negligible level.

The signal is then fed to a precision rectifier, or "super diode". As discussed above, this circuit takes advantage of the high open-loop gain of an operational amplifier to wash out the non-linearity of a standard silicon signal diode placed in the feedback loop. The result is a wide dynamic range high linearity precision rectifier. (This is in sharp contrast to the limited dynamic range and poor linearity of the conventional bridge rectifier circuit employed in the original TT-1A AC meter as per Figure 14.)

This rectified signal voltage is then passed through a lowpass filter which integrates it, thereby converting it into a smooth DC voltage. A final operational amplifier is then used to buffer and amplify this DC voltage so that it can drive the TT-1A meter. Since the necessary signal rectification has already been done with the new added circuitry, this meter output can be applied to the TT-1A meter DC (rather than AC) input terminals for improved accuracy (i.e., bypassing the non-linear internal bridge rectifier circuit).

A regulated power supply is also included to provide the necessary +/-12 VDC operating voltages. Note that this power supply must be isolated from all other TT-1A circuitry (including the TT-1A chassis) since the tube under test output circuitry is operated at a high voltage.

To explain the rationale for the unusual plate load resistance (R_L) value of 20.833 ohms, keep in mind that the TT-1A is capable of measuring tube transconductance values as high as 24,000 micromhos. For the sake of ease of calibration and simplified computation, I arbitrarily selected R_L so that a tube with a transconductance of 24,000 micromhos would have a voltage gain (A_V) equal to precisely 0.5. Restating the voltage gain equation (2) from above:

$$A_V = G_m \times R_L \tag{4}$$

Rearranging this equation to solve for R_L:

$$R_L = A_V/G_m \tag{5}$$

Solving equation (5) for Av = 0.5 and G_m = 24,000 micromhos (or 0.024 mhos for unit consistency) as discussed above, we obtain R_L = 20.833 ohms. Although this is not a standard 1% value, a 37.4 ohm 1% resistor in parallel with a 47.0 ohm 1% resistor provides a composite resistance of 20.827 ohms, which is a very close match (and well within the 1% tolerance).

Of course, there is nothing magic about using 20.833 ohms. However, this value is convenient for calibration purposes.

The step attenuator in the functional block diagram of Figure 15 also requires explanation. Essentially, this is a selectable resistive voltage divider that appropriately adjusts the meter drive level, replacing the front-panel Meter control that is now inoperative for transconductance measurements. Unlike the Meter control, however, this step attenuator is very precise.

To set up the background for this, the reader will recall that the TT-1A G_m meter scale is calibrated in “Proportional Micromhos” on a scale of 0-3000. Actual transconductance is then found by multiplying the indicated Proportional Micromhos by the appropriate multiplier (X=1 through X=8) appearing in the far right “Remarks” column on the roll-chart data line for the tube under test. For a 6AU6, for example, the multiplier is X=2, so a Proportional Micromhos meter reading of 1900 would be multiplied by 2 for a true transconductance of 3800 micromhos.

The X multiplier depends on the signal input level (as set by the Signal selector switch), the tube transconductance, and especially the front-panel Meter control (for an unmodified TT-1A). It appears that this X multiplier was primarily selected to result in a meter deflection in the upper portion of the meter scale for the tube under test. A upper-end meter scale reading serves both to improve meter indication readability and to avoid the excessive meter non-linearity (caused by the meter rectifier diodes) present in the lower portion of the meter scale.

A design goal of the improved TT-1B was to preserve the TT-1A meter reading convention (i.e., to keep the same 0-3000 Proportional Micromhos scale and retain the same X multiplier convention). Although the 0-3000 scaling goal was met, the required X multipliers were different than the originals for reasons discussed in subsequent Sections.

Since the front-panel Meter control is inoperative for transconductance measurements in the TT-1B, the new "X Select" precision step attenuator switch is substituted. As per Figure 16, this attenuator selects the X multiplier values as appropriate from 1 through 8, with an additional X=0.5 value included primarily for calibration purposes.

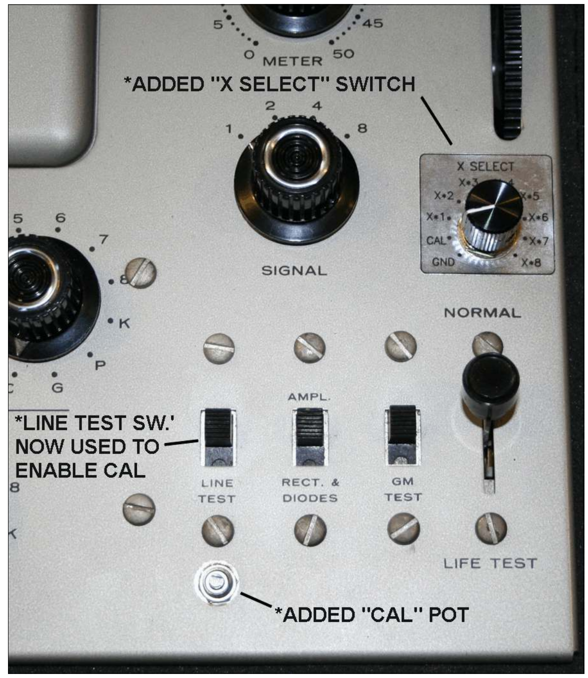


Figure 16 - Added TT-1B Front-Panel Controls (note that Line Test switch is now the Calibration Enable switch)

This additional X=0.5 multiplier is included to facilitate a new TT-1B front-panel CAL (calibration) procedure. Readers familiar with the original TT-1A calibration procedure will recall that it is a lengthy and tedious process that requires removing the unit from its case in order to access internal adjustments. The TT-1B, in sharp contrast, includes an added front-panel CAL adjustment (see Figure 16) that allows the new calibration procedure to be conveniently done in just a few seconds without the need for removing the unit from its case. Also as per Figure 16, the Line Test switch, which is no longer needed for its original functional, has been converted into a calibration enable switch. Figure 17 shows the rear view of the added X Select switch.

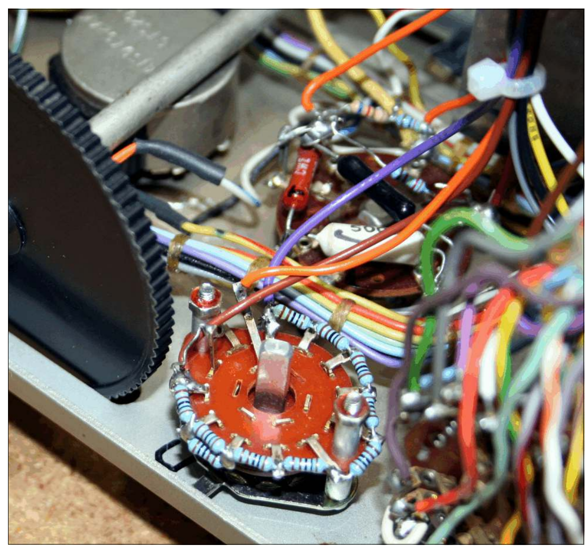


Figure 17 - Added X Select Switch Rear View (note attenuator resistors mounted directly on switch)

A more detailed explanation of the added X Select switch and CAL function is provided in Sections VII and VIII and Appendix B.

E. ENHANCED TRANSCONDUCTANCE TEST CIRCUIT DETAILED DESCRIPTION

With the above functional description and Figure 15 in mind, the following paragraphs present a detailed circuit description. Referring to the detailed signal amplifier/detector and meter driver schematic of Figure 19, R5 is the 20.833 ohm tube plate load resistor (R_L in Figure 13). The 5.030 kHz signal voltage developed across this load resistor is then passed on to relay K1 (Op/Cal), which allows the 200 mVPP signal oscillator calibration voltage to be substituted for the tube output signal voltage during calibration. R1 is included to raise the tube output circuit drive resistance so that it is close to the 250 ohm calibration signal source output resistance. (Equalizing these drive resistances equalizes the subsequent differential amplifier gain in both the Op and Cal modes.)

The signal is then applied to an RC highpass filter comprising C1, C2, and R6 that suppresses 50/60 Hz power supply ripple, followed by high-gain differential amplifier U1:A. This amplified signal is then passed through another RC highpass filter comprising C3 and the nine X Select step attenuator resistors (8000 ohms total). This highpass filter provides further 50/60 Hz power supply ripple rejection. The net voltage gain (Av) through the differential amplifier and two highpass filters is close to 45.

The X Select step attenuator output is then passed through op amp U1:B with a voltage gain of close to 33. This amplified signal is then passed through another highpass filter comprising C4, R8, and R9 that provides still more 50/60 Hz power supply ripple rejection. Shunt capacitor C5 provides high frequency noise rejection.

Op amps U1:C, U1:D, plus diodes D1, D2, and associated components comprise a precision half-wave rectifier (or "super diode") that provides highly linear 5.030 kHz signal rectification over a wide dynamic range. U1:C serves as the active rectifier while U1:D serves as a high input impedance unity-gain buffer amplifier. This circuit provides linearity and dynamic range that is superior to the TT-1A AC meter rectifier by orders of magnitude. Figure 18 illustrates the output waveform (at U1-14) when the TT-1A is in the CAL mode.

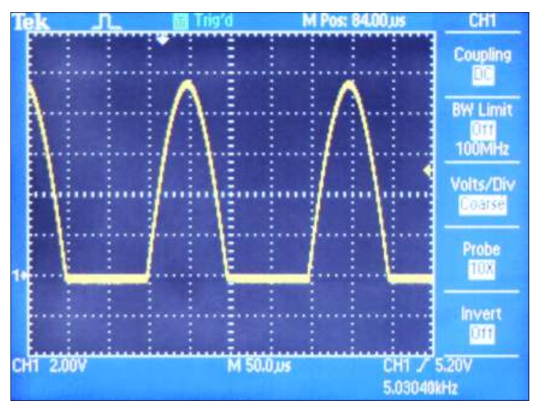
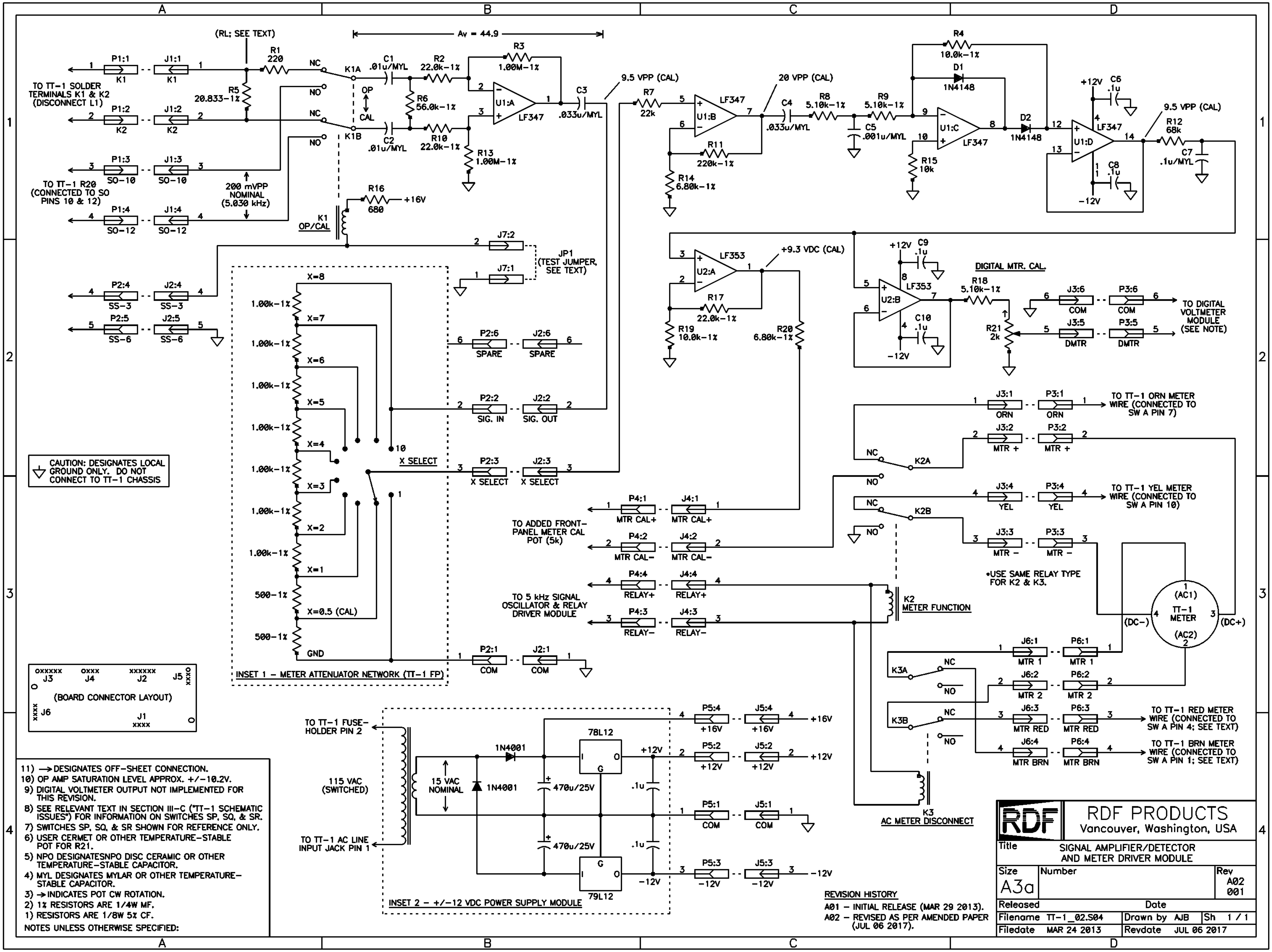


Figure 18 - Precision Half-Wave Rectifier Output Waveform((U1-14; CAL mode)

The half-wave rectified signal is then passed through RC lowpass filter R12/C7 to provide a smooth (averaged) proportional DC voltage. This DC voltage is then applied to meter driver op amp U2:A, which has a voltage gain (Av) of 3.2. The U2:A output is then applied through current limiting resistor R20, the added front-panel 5k meter CAL pot, and then to the TT-1A meter through DPDT relay K2. R20 (6.80k) is sufficiently large in value to limit meter drive current to a safe value (i.e., that will prevent damage to the meter).

Note that this meter drive voltage is applied to the TT-1A DC meter terminals (pins 3 and 4) rather than its AC meter terminals (pins 1 and 2 as per the original design). To explain, it would defeat the purpose of designing an elaborate precision high-linearity signal rectifier and then drive the TT-1 meter AC input with its non-linear rectifier diodes.

Figure 19 needs to be replaced ?



(RL; SEE TEXT)
Av = 44.9
TO TT-1 SOLDER TERMINALS K1 & K2 (DISCONNECT L1)
TO TT-1 R20 (CONNECTED TO SO PINS 10 & 12)
200 mVPP NOMINAL (5.030 kHz)
P1:1 K1
J1:1 K1
P1:2 K2
J1:2 K2
P1:3 SO-10
J1:3 SO-10
P1:4 SO-12
J1:4 SO-12
P2:4 SS-3
J2:4 SS-3
P2:5 SS-6
J2:5 SS-6
R1 220
R5 20.833-1%
K1A
K1B
OP
CAL
K1 OP/CAL
R16 680
+16V
C1 .01u/MYL
C2 .01u/MYL
R2 22.0k-1%
R3 1.00M-1%
R6 56.0k-1%
R10 22.0k-1%
R13 1.00M-1%
U1:A LF347
C3 .033u/MYL
9.5 VPP (CAL)
R7 22k
U1:B LF347
R11 220k-1%
R14 6.80k-1%
20 VPP (CAL)
C4 .033u/MYL
R8 5.10k-1%
R9 5.10k-1%
C5 .001u/MYL
U1:C LF347
R15 10k
R4 10.0k-1%
D1 1N4148
D2 1N4148
U1:D LF347
+12V
-12V
C6 .1u
C8 .1u
R12 68k
C7 .1u/MYL
J7:2
J7:1
JP1 (TEST JUMPER, SEE TEXT)
U2:A LF353
+9.3 VDC (CAL)
R17 22.0k-1%
R19 10.0k-1%
R20 6.80k-1%
C9 .1u
C10 .1u
U2:B LF353
DIGITAL MTR. CAL.
R18 5.10k-1%
R21 2k
J3:6 COM
P3:6 COM
J3:5 DMTR
P3:5 DMTR
TO DIGITAL VOLTMETER MODULE (SEE NOTE)
X=8
X=7
X=6
X=5
X=4
X=3
X=2
X=1
X=0.5 (CAL)
GND
1.00k-1%
500-1%
X SELECT
INSET 1 - METER ATTENUATOR NETWORK (TT-1 FP)
P2:6 SPARE
J2:6 SPARE
P2:2 SIG. IN
J2:2 SIG. OUT
P2:3 X SELECT
J2:3 X SELECT
P2:1 COM
J2:1 COM
CAUTION: DESIGNATES LOCAL GROUND ONLY. DO NOT CONNECT TO TT-1 CHASSIS
J3:1 ORN
P3:1 ORN
TO TT-1 ORN METER WIRE (CONNECTED TO SW A PIN 7)
J3:2 MTR +
P3:2 MTR +
K2A
NC
NO
J3:4 YEL
P3:4 YEL
TO TT-1 YEL METER WIRE (CONNECTED TO SW A PIN 10)
K2B
J3:3 MTR -
P3:3 MTR -
•USE SAME RELAY TYPE FOR K2 & K3.
P4:1 MTR CAL+
J4:1 MTR CAL+
P4:2 MTR CAL-
J4:2 MTR CAL-
TO ADDED FRONT-PANEL METER CAL POT (5k)
P4:4 RELAY+
J4:4 RELAY+
P4:3 RELAY-
J4:3 RELAY-
TO 5 kHz SIGNAL OSCILLATOR & RELAY DRIVER MODULE
K2 METER FUNCTION
(AC1)
TT-1 METER
(AC2)
(DC-)
(DC+)
J6:1 MTR 1
P6:1 MTR 1
J6:2 MTR 2
P6:2 MTR 2
J6:3 MTR RED
P6:3 MTR RED
J6:4 MTR BRN
P6:4 MTR BRN
K3A
K3B
TO TT-1 RED METER WIRE (CONNECTED TO SW A PIN 4; SEE TEXT)
TO TT-1 BRN METER WIRE (CONNECTED TO SW A PIN 1; SEE TEXT)
K3 AC METER DISCONNECT
oxxxxx J3
oxxx J4
xxxxxx J2
J5 oxxx
xxxx J6
J1 xxxx
(BOARD CONNECTOR LAYOUT)
TO TT-1 FUSE-HOLDER PIN 2
115 VAC (SWITCHED)
TO TT-1 AC LINE INPUT JACK PIN 1
15 VAC NOMINAL
1N4001
1N4001
470u/25V
470u/25V
78L12
79L12
.1u
.1u
-12V
P5:4 +16V
J5:4 +16V
P5:2 +12V
J5:2 +12V
P5:1 COM
J5:1 COM
P5:3 -12V
J5:3 -12V
INSET 2 - +/-12 VDC POWER SUPPLY MODULE
11) → DESIGNATES OFF-SHEET CONNECTION.
10) OP AMP SATURATION LEVEL APPROX. +/-10.2V.
9) DIGITAL VOLTMETER OUTPUT NOT IMPLEMENTED FOR THIS REVISION.
8) SEE RELEVANT TEXT IN SECTION III-C ("TT-1 SCHEMATIC ISSUES") FOR INFORMATION ON SWITCHES SP, SQ, & SR.
7) SWITCHES SP, SQ, & SR SHOWN FOR REFERENCE ONLY.
6) USER CERMET OR OTHER TEMPERATURE-STABLE POT FOR R21.
5) NPO DESIGNATESNPO DISC CERAMIC OR OTHER TEMPERATURE-STABLE CAPACITOR.
4) MYL DESIGNATES MYLAR OR OTHER TEMPERATURE-STABLE CAPACITOR.
3) → INDICATES POT CW ROTATION.
2) 1% RESISTORS ARE 1/4W MF.
1) RESISTORS ARE 1/8W 5% CF.
NOTES UNLESS OTHERWISE SPECIFIED:
REVISION HISTORY
A01 - INITIAL RELEASE (MAR 29 2013).
A02 - REVISED AS PER AMENDED PAPER (JUL 06 2017).
RDF
RDF PRODUCTS
Vancouver, Washington, USA
Title SIGNAL AMPLIFIER/DETECTOR AND METER DRIVER MODULE
Size A3a
Number
Rev A02 001
Released
Date
Filename TT-1_02.S04
Drawn by AJB
Sh 1 / 1
Filedate MAR 24 2013
Revdate JUL 06 2017

F ure 19

A complication that arises using DC meter terminals, however, is that other TT-1A circuitry is still connected to it with the result that it will not function properly. This problem is averted using K2, which isolates the DC meter terminals away from other TT-1A circuitry during transconductance measurements. Referring to Figure 19, when transconductance is not being measured, K2 connects the DC meter terminals to the TT-1A circuitry as per the original TT-1A schematic. When the TT-1A Gm Test switch is pulled down (for transconductance measurement), however, K2 activates, disconnects the DC meter terminals from this other TT-1A circuitry, and switches these terminals exclusively to the meter driver amplifier output (U2-1, through R20 and the added 5k front-panel CAL pot).

An additional complication is that other TT-1A circuitry is still connected to the AC meter terminals during transconductance measurements with the result that undesired voltages can be applied to the meter via this path. Although the AC meter is no longer used for transconductance measurements in the new design, the wires driving the AC meter cannot just be cut since the AC meter is also used in the TT-1A tube leakage tests. To solve this problem, an additional relay (K3) is employed to disconnect the AC meter terminals from the other TT-1A circuitry. K3 is enabled in tandem with K2.

An additional meter driver op amp (U2:B) is employed as a digital meter driver. Although this feature has not yet been implemented, a possible further design improvement for the TT-1B is the inclusion of a digital voltmeter circuit to indicate transconductance on a numeric display. The digital meter driver would facilitate this improvement if it is pursued.

Op amps U1 and U2 require regulated +/-12 VDC supplies for this application. This was done using an available small AC-to-DC power adaptor (which is inexpensive and readily available from many sources). The only component of this adaptor I used was the power transformer and enclosure. Any small power transformer with a 15-18 VAC secondary and capable of 200 mA AC current output will suffice.

As per Figure 19, the transformer secondary drives two half-wave rectifier circuits (one a positive DC output and the other a negative DC output). These DC outputs are filtered by 470 uF/25V capacitors and then fed to 78L12/79L12 (positive/negative) 12 volt regulators. Note: 78/79-series 3-terminal regulators require a minimum input overhead voltage of 3.0 volts for conservative design practice to guarantee that they do not fall out of regulation. The unregulated positive output (nominally +16 VDC) is also brought out to power the Op/Cal relay (K1) coil.

K1, K2, and K3 are all miniature DPDT relays with 12 VDC coils. I recommend relays with required coil drive currents of 20 mA or less. Since K2 and K3 are driven in parallel, they should be of the same type. Note: Some relays have polarized coils and will only activate when current flows through these coils in the proper direction.

Although many different op amps can be used at U1 and U2, substitutions should be done advisedly (particularly for U1). The LF347 and LF353 used for U1 and U2, respectively, are high input impedance JFET input op amps with low input currents and DC offsets. Since U1:A and U1:B are configured as high-gain amplifiers, these good DC characteristics are important.

The most important performance attribute of U1:C (the LF347 op amp section used for the active rectifier) is high gain and wide bandwidth. Active rectifiers require high open-loop bandwidth at the operating frequency for good precision. With a test signal frequency of 5.030 kHz, a reasonably high gain/high speed op amp like the LF347 is required for good performance. Since the LF347 is inexpensive, readily available, and verified for use in this circuit, I don't recommend a substitution.

F. ADJUSTMENT AND CALIBRATION

The 4-position internal "Calibrate" switch (located inside the unit underneath the roll-chart) is no longer used for TT-1B calibration. This switch must be permanently set to its Operate position for both calibration and operation.

Since the circuit employs precision components, very little adjustment is necessary. Once the circuit wiring has been debugged and verified and the DC supply voltages have been checked, the best way to test and adjust the circuit is to conduct the calibration procedure. (Note that this requires that the grid bias voltage regulator and signal oscillator enhancements presented earlier in this paper have been satisfactorily completed.) Set the new X Select switch to its CAL position and then pull down the Line Test/Cal switch to enable the calibration mode. This simultaneously activates the signal oscillator module and K1 Op/Cal relay. For hands-free convenience, this can alternatively be accomplished by installing the test jumpers JP1 on both the signal oscillator and signal processing modules. Be sure that no tube is inserted in any of the TT-1B test sockets during this calibration procedure.

After first setting R10 (the signal oscillator module output amplitude adjustment pot) for minimum output, probe the first 9.5 VPP calibration test point (at C3 on the signal processor module) with an oscilloscope and verify the presence of a clean 5.030 kHz sinusoidal waveform. Adjust R10 for an indicated amplitude of 9.5 VPP. Next, probe the 20 VPP calibration test point (at U1-7 on the signal processor module) and readjust R10 for an indicated waveform amplitude of 20 VPP. *Confirm that this waveform is a clean 5.030 kHz sinusoid with no clipping.*

Next, probe the second 9.5 VPP calibration test point at U1-14 on the signal processor module) and confirm the presence of a half-wave rectified sinusoidal waveform close to 9.5 VPP very similar in appearance to that of Figure 18. Once done, probe the +9.3 VDC test point at U2-1 and confirm that the indicated voltage is very close to +9.3 VDC.

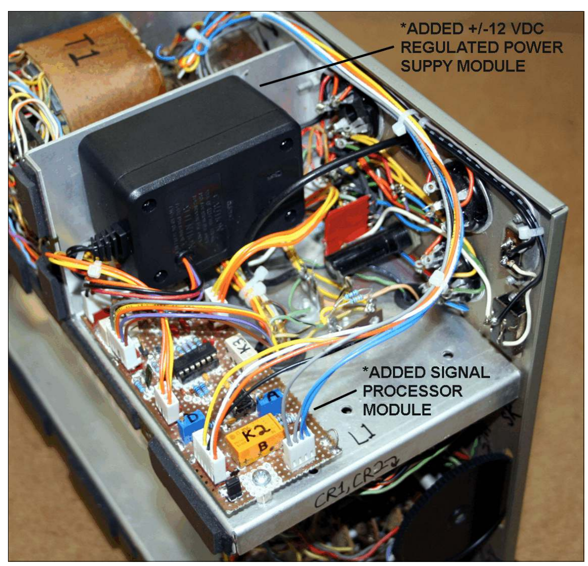


Figure 20 - Added Signal Processor and Regulated Power Supply Modules

Next, observe the TT-1B meter and adjust the added front-panel CAL pot for a full-scale indication. Finally, release the Line Test/Cal switch (or pull the jumpers) to disable the calibration mode and confirm that the TT-1B meter drops to zero.

As a final check, again pull down the Line

Test/Cal switch (or reinstall the jumpers) and confirm that the 5.030 kHz signal as measured across the TT-1B 250 ohm signal attenuator resistor (R20; connected between pins 10 and 12 of the TT-1B Signal switch SO) is near its nominal 200 mVPP value.

The added signal processor and regulated power supply modules are illustrated in the photo of Figure 20. Note that choke L1 and the large oil-filled capacitor (C9/C10/C11) have been removed to make room for the signal processor module. (As discussed in VR-004, the oil-filled capacitor is unnecessary.)

G. REQUIRED CHANGES TO THE TT-1A

As per the Figure 19 schematic, the new signal processor module must be connected to the new 5.030 kHz signal oscillator module and +/-12 VDC power supply. Additionally (and also as per the Figure 19 schematic), it must be connected to the added X-Select meter attenuator network and the added front-panel calibration adjustment pot (see Figures 16 and 17).

Also, it must also be appropriately wired into the TT-1A. This wiring is discussed below and also illustrated in the marked-up TT-1A schematics of Figures 21 and 22 for clarity. Referring to Figures 19, 21 and 22, these changes are summarized as follows:

a. P1:1/P1:2 (K1/K2) must be wired to TT-1A solder terminals K1 and K2, respectively, and plate choke L1 removed as per Figure 21. This change effectively substitutes the new 20.833 ohm plate load resistor R5 for the discarded plate choke to facilitate precision transconductance measurements with a low-value well-defined plate load resistance.

b. P2:4/P2:5 (SS-3/SS-6) must be wired to TT-1A Line Test switch terminals SS-3 and SS-4, respectively, as per Figure 21. (SS is actually used as a calibration enable switch in the modified design - this change allows SS to enable the calibration mode.)

c. P1:3/P1:4 (SO-10/SO-12) must be wired to TT-1A Signal switch terminals SO-10 and SO-12, respectively, as per Figure 21. This change applies the 200 mVPP calibration signal voltage to the new signal processor module.

d. P3 must be wired into the TT-1A meter circuitry as illustrated in Figure 22. This change selects the DC meter for transconductance measurements. As discussed, with the improved circuit the AC meter is no longer used for transconductance measurements due to its non-linearities. Instead the highly accurate DC meter is used. This is accomplished using DPDT relay K2.

e. P6 must be wired into the TT-1A meter circuitry as illustrated in Figure 22. This change disconnects the AC meter from all TT-1A circuitry during transconductance measurements. As discussed, if the AC meter is not completely isolated during transconductance measurements, undesired voltages can be applied to the meter via this path and impair the accuracy. This is accomplished using DPDT relay K3.

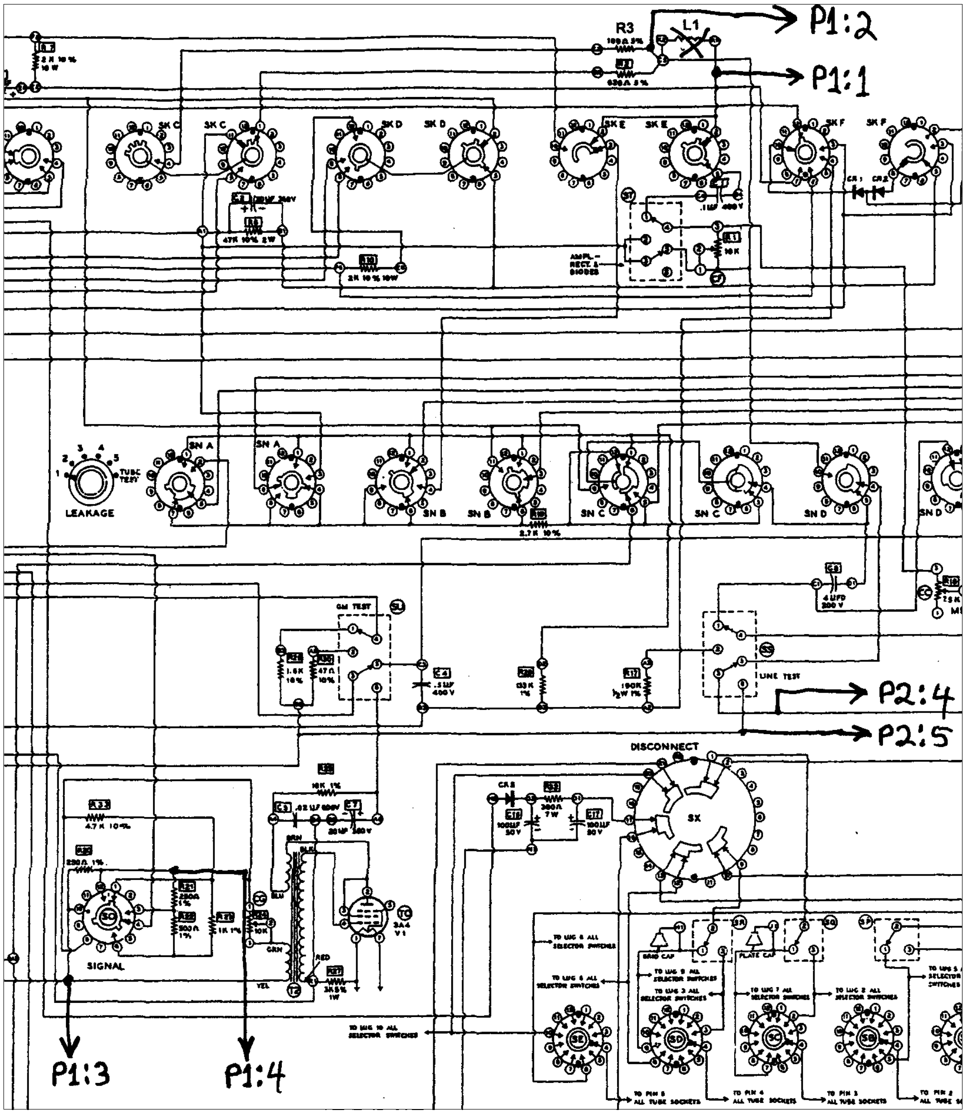


Figure 21 - Wiring the Signal Processor Module into the TT-1A (1)

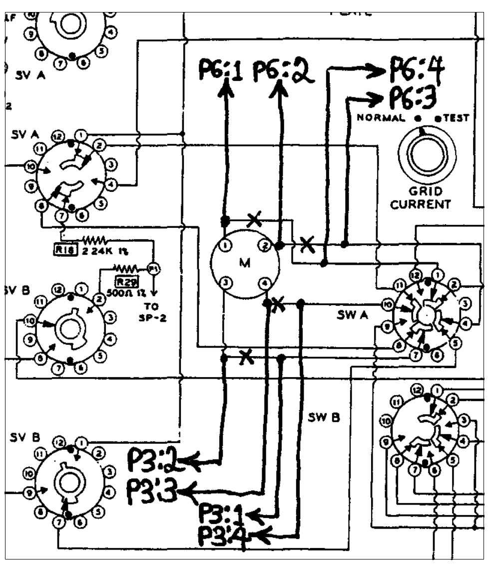


Figure 22 - Wiring the Signal Processor Module into the TT-1A (2)

H. MISCELLANEOUS ISSUES

1. OSCILLOSCOPE MEASUREMENT DIFFICULTIES

When testing the new signal processor module with the oscilloscope, I noticed that I was able to obtain meaningful measurements only when the TT-1B was plugged into the AC power line with a particular polarity. If the AC power line connection was reversed, noise or hum pickup corrupted the oscilloscope display for some reason.

This seems to be only a measurement anomaly associated with the oscilloscope. Since the TT-1B functions equally satisfactorily with the AC power connected either way, I did not pursue this. Even so, readers should be aware of this issue.

2. X SELECT SWITCH ARTWORK AND LASER LABEL

The X Select switch label artwork is presented in Figure 23 for the reader's convenience. I have also placed this artwork in electronic file format on the N6DC vintage radio website (file x-select_02.tif). This artwork was created in Tango PCB (my printed circuit board layout program). Note that this artwork was created for a switch rotating in 30° increments with a bushing diameter suitable for a 3/8" mounting hole and a ½" diameter knob. The label diameter is 1.2" for reference.

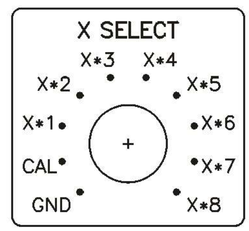


Figure 23 - X Select Switch Label Art

I printed the label (actually a sheet of 48 labels) on a blank 8.5" x 11" metallic silver label sheet. These label sheets are designed for laser printers and are available from Paper Direct (www.paperdirect.com; P/N KL1004). After the sheet has been printed, the label(s) can be cut out and the protective backing removed to expose the adhesive surface. Since the label is difficult to position correctly, having the extra labels allows repeated tries.

SECTION VII - OPERATING THE TT-1B

The TT-1B operates mostly the same as the TT-1A. The non-transconductance tests (e.g., the leakage, grid current, diode, rectifier, voltage regulator, thyratron, and magic-eye tests) are all conducted normally as per the Heathkit TT-1 operational manual. The only procedural change for these tests is that the Set Line control is now used to adjust the tube filament voltage to its correct value using the external AC DVM as discussed in Section IV-D. The only exception to this procedural change is that the Set Line control is used in its standard fashion for testing voltage regulator tubes (which, of course, do not have filaments).

Although TT-1B transconductance testing is similar to that of the TT-1A, there are important differences. These procedural changes are clarified in the following summary transconductance test procedure:

1. Place the TT-1B in its normal horizontal position on a flat table or bench. (If the TT-1B is set vertically, its meter will be less accurate. Also, some tubes are were not designed for horizontal mounting.)

2. With no AC power yet applied, verify that the meter indicates a precise zero reading. Adjust the meter zero-set screw as required.

3. *With no tube yet inserted*, set up the TT-1B for a transconductance test. *Read the cautionary note below regarding the X Select, SIG and Bias setups.* Since the Meter switch is no longer used for Gm testing, it can be left at any setting with no effect.

4. To verify calibration, power-up the TT-1B (with no tube yet inserted), set the X Select switch (see Figure 16) to its CAL position and then pull down the Line Test switch. This should result in a precise meter full-scale indication (i.e., 3000 Proportional Micromhos). If this indication is in error, correct it using the added calibration adjustment pot as per Figure 16. No warm-up time is required for calibration.

   Note: This calibration adjustment should mostly be "set-and-forget", not requiring frequent readjustment. If any calibration "drift" is noticed (i.e., if the calibration pot requires frequent readjustment), this is an indication that certain critical components in the new signal oscillator and signal processor modules may not be temperature- or age-stable. As per the schematics and the relevant text above, critical resistors should be 1% metal film and critical capacitors should be Mylars or NPO disc ceramics.

5. Set the SIG and added X Select switches to their appropriate settings.

CAUTION

The SIG and X Multiplier settings specified in the roll-chart or tube list are unsuitable for the TT-1B. The correct settings are found using the operating software presented and discussed in detail below. The only roll-chart/tube list settings used are the "Selectors" switch settings, plus the K1, P1, and G1 settings and plate/grid caps (if used).

6. Plug the tube to be tested into the appropriate test socket and wait at least one minute for the tube filament to reach its full operating temperature. Monitor the tube filament voltage with an AC DVM as per Section IV-D, adjusting the Line Set control as required for a precise filament voltage indication (e.g., 6.3 VAC for a 6AU6). Note that the indicated filament voltage rises as the filament warms up, so be sure to readjust the Line Set control as required until the filament voltage fully stabilizes.

7. Pull down the G_m test switch and read the transconductance indication from the Proportional Micromhos scale in the normal fashion. If the meter indication continues to slowly rise, this means that the filament has not yet reached its full operating temperature.

8. The correct Reject Point ("R.P.") value is determined by the TT-1B_01.BAS computer program rather than that specified in the roll-chart/tube list. As per the Heathkit TT-1 operational manual, this reject point is based on an EIA/RETMA standard that arbitrarily defines the tube reject point as 65% of the nominal transconductance value for a factory-fresh tube. However, as the manual further states, this value is application dependent and thus subject to interpretation.

9. Determining actual tube transconductance is done by multiplying the indicated 0-3000 micromho proportional transconductance by a suitable multiplier factor provided by the computer program rather than the X multiplier specified in the roll-chart/tube list. In most cases, this new "P-Multiplier" will not be the same value as the X Select switch setting. Additionally, this P-Multiplier can be either an integer or fractional number, and can range from values less than 1 to greater than 8.

10. The Leakage, Grid Current, and Life Tests can be done in the normal fashion as per the manual.

To conclude this Section with a slightly off-topic subject, vacuum tube rectifiers and signal diodes (e.g., the 5Y3 and 6AL5, respectively) can often be replaced with inexpensive modern solid-state components offering superior performance. Although some vintage radio enthusiasts may object to this practice based on "authenticity", this must be weighed against the many advantages of solid-state diode substitution. Similarly, modern Zener diodes can be substituted for voltage regulator tubes. This will usually require a series string to handle the higher voltages.

My philosophy is that the authenticity issue is greatly outweighed by the significant benefits which include component availability, cost, reliability, and the reduced load burden placed on irreplaceable radio power transformers. This topic is covered in much greater depth in VR-003 for interested readers.

At the end of the day, each vintage radio enthusiast must decide for himself where he wants to draw the line on this authenticity issue. I've drawn my line to allow substitution of any passive component (e.g., resistors, capacitors, diodes, etc). I would not attempt to replace amplifier tubes, however, with transistors or ICs unless there was no other way to complete the reconditioning project.

SECTION VIII - DETERMINING THE X SELECT AND SIG SWITCH SETTINGS

A. OVERVIEW

During the course of the TT-1B redesign it was my expectation that the added X Select switch would simply be set to whatever X multiplier was specified in the roll-chart/tube list. It was my further expectation that this selected X value would both a) result in a proper, mid- to upper-scale meter deflection; and b) would, as is the case in the original TT-1A design, correctly multiply the Proportional Micromho meter reading to the actual tube transconductance.

To my surprise, neither of these two expectations turned out to be the case for the reasons discussed in depth below. As also discussed below, a computer program had to be written to compute the correct X Select and SIG switch settings as well as the correct multiplier necessary to multiply the meter Proportional Micromho reading to actual tube transconductance.

B. OFF-SCALE/LOW-SCALE READINGS

Throughout the course of this TT-1B project, I used a known good 6AU6 pentode as the tube under test to validate the various transconductance equations presented in this paper with good results. When I then substituted a 6BA6 (a very similar pentode), I was stunned to see an off-scale meter reading using the roll-chart specified X multiplier (which was different than that specified for the 6AU6). After some investigation, I determined that this unexpected result was caused by the afore-mentioned vagaries associated with the TT-1A Meter control setting.

As per that discussion, the Heathkit engineers ultimately had to rely on this Meter control to compensate for the various imprecisions inherent in the original design (i.e., the ill-defined plate load resistance R_L and the AC meter non-linearities). This Meter control setting (on a scale of 0-50) appears to be empirically derived and was never intended to be used as part of a precise transconductance equation. Since this Meter control setting was not constrained by any precise computational structure, the Heathkit engineers had the latitude to choose from among more than one X multiplier numbers and then compensate for this by specifying an appropriate Meter control setting. Suffice it to say that they placed very heavy reliance on this Meter control to achieve the desired results.

As a case in point, the roll-chart specifies a 6BA6 X multiplier of 3 (X3) and a Meter control setting of 33. However, they could also have specified an X multiplier of 2 (X2) and some higher Meter control setting.

In contrast, the TT-1B X Select switch (which effectively replaces the TT-1A Meter control) must be set to a very specific value (0.5-8) that is consistent with the constraints imposed by a precise transconductance equation. Full-scale meter transconductance is computed as follows:

$$G_m = (3{,}000 \times SIG)/X \quad (6)$$

where:

Gm is the meter full-scale transconductance (in micromhos),
SIG is the Signal selector switch (signal oscillator) attenuator setting (1, 2, 4, or 8),
X is the X Select step attenuator setting (0.5, 1, 2, 3, 4, 5, 6, 7, or 8).

As per equation (6), the only available degrees of freedom are the SIG (signal drive level) and X Select settings. If we go with the SIG setting specified on the roll-chart, then for most tubes there is only one X multiplier value that will result in a factory-fresh tube Gm indication in the top half of the meter scale (preferred for accuracy and readability).

To avoid belaboring this issue with additional explanation and tedious technical detail, *suffice it to say that the original TT-1A roll-chart X multiplier settings are not suitable for the TT-1B for most tubes. A different X multiplier (X Select) setting must then determined in these cases.*

A further (although lesser) complication that emerged is that *in many cases the new X Select value is not an appropriate multiplier for computing actual tube transconductance* (although it is fully satisfactory for the TT-1B's primary Gm pass/fail test). The reader will recall that the sole purpose of the published X multiplier value in the original TT-1A design was to allow computation of true transconductance by multiplying the Proportional Micromhos reading by X.

Although this shortcoming does not impair the TT-1B's ability to perform its primary Gm pass/fail test, most users would prefer to have a straightforward means to convert the Proportional Micromhos meter indication to actual transconductance. Fortunately, the correct "P-Multiplier" can be easily computed.

C. TT-1_01.BAS COMPUTER PROGRAM

***UPGRADE FOR VR-005B: See Appendix H for modernized Windows software program.**

Although only modest math skill is necessary to compute more appropriate X Select, P-Multiplier, and revised R.P. values, it is not a task that we would want to have to do on a recurring basis every time we test a tube. To remedy this inconvenience, I wrote an easy-to-use computer program (TT-1B_01.BAS) that makes this task simple and straightforward. See Appendix E for a full source-code listing and other important information.

The only required initial user input for this program is the tube factory-fresh transconductance value. The program then computes and displays the recommended initial X Select and SIG switch settings, the R.P. value, the Proportional Micromho multiplier, and other useful information.

Operation of this program is best explained by example using a 6BA6 remote-cutoff pentode. After starting the program and following the instructions on the opening screen, the following prompt appears (user responses are shown in bold-faced type):

```
*ENTER FACTORY-FRESH TUBE Gm (375-48,000 uMHOs)  4300
```

This Gm number (4300) came out of the GE “Essential Characteristics” tube manual (see reference 6), but obtaining this number is a topic for the fuller discussion that follows. Also note that the TT-1B permits a much wider Gm measurement range than its predecessor TT-1A.

Once this information is entered, the program computes and displays the following eight output lines:

```
**FACTORY-FRESH TUBE Gm =  4300
**65% RETMA TUBE REJECT POINT Gm =  2795
**X SELECT SWITCH SETTING =  5
**SIG SWITCH SETTING =  8
**METER FULL-SCALE Gm =  4800
**TUBE REJECT POINT METER INDICATION =  1747
**FACTORY-FRESH TUBE Gm % FULL-SCALE METER INDICATION =  90
**PROPORTIONAL uMHO MULTIPLIER =  1.600
```

For a fuller explanation of this output information:

Line 1 - This is just a restatement of the user-input factory-fresh tube Gm for convenience of reference (4300 for the 6BA6).

Line 2 - This is the old EIA/RETMA tube transconductance pass/fail criterion which was defined (somewhat arbitrarily) as 65% of the factory-fresh tube Gm. For the 6BA6, this value is 4300 x 0.65 = 2795. This is presented for convenience of reference.

Line 3 - This is the computed recommended initial setting of the TT-1B added X Select switch.

Line 4 - This is the computed recommended initial setting of the TT-1B SIG switch.

Line 5 - This is the TT-1B computed meter full-scale Gm value for the specified X Select and SIG switch settings. It is computed by multiplying the SIG switch setting by 3000 and dividing the result by the X Select setting. For the 6BA6, this value is (3000 x 8)/5 = 4800.

Line 6 - This is the computed tube reject point meter indication (on the 0-3000 Proportional Micromho scale) for the specified X Select and SIG switch settings. It is computed by dividing the full-scale Proportional Micromho value of 3000 by the meter full-scale Gm value computed in Line 5 and then multiplying that quotient by the 65% RETMA tube reject point from Line 2. For the 6BA6, this value is (3000/4800) x 2795 = 1747 (rounded to the nearest whole number). This 1747 number is thus the 6BA6 RETMA pass/fail threshold on the TT-1B 0-3000 Proportional Micromho scale.

Line 7 - This is the percentage full-scale meter reading that would be obtained with a factory-fresh tube. This percentage is computed by dividing the factory-fresh tube Gm by the meter full-scale Gm reading computed in Line 5 and multiplying that quotient by 100. For the 6BA6, this value is (4300/4800) x 100 = 90% (rounded to the nearest whole number).

Line 8 - This is the computed multiplier factor applied to the meter 0-3000 Proportional Micromho scale to convert the Proportional Micromho reading to true Gm (i.e., the aforementioned "P-Multiplier"), and is computed by dividing the SIG switch setting by the X Select switch setting. For the 6BA6, this value is 8/5 = 1.600. If we multiply the reject point Gm on the 0-3000 Proportional Micromho scale from Line 6 by the computed 1.600 multiplier factor, the result is 1747 x 1.600 = 2795 (which correctly matches the 2795 R.P. value given in Line 2). Once again, results are rounded to nearest whole numbers for better presentation.

A succinct list of the above equations is presented in Appendix A.

Focusing once again on the percentage full-scale meter reading discussed in Line 7, there is an issue associated with this number that warrants further discussion. Ideally, we would prefer that the maximum meter reading for a factory-fresh tube be in the upper portion of the meter display for best accuracy and presentation (i.e., 60-90% full-scale). The 90% full-scale indication for a factory-fresh 6BA6 as per Line 7 is thus very favorable. As written, the program finds X Select and SIG switch settings that will always result in factory-fresh tube full-scale indications of 50-100%. Although most tubes will fall comfortably inside this range, there are also some that will fall uncomfortably close to the 50/100% boundaries.

As a case in point, if we run the program for a 6AU6 (a sharp-cutoff pentode very similar to the 6BA6), we obtain the following results (for an entered tube transconductance of 3900 micromhos):

```
**FACTORY-FRESH TUBE Gm = 4000
**65% RETMA TUBE REJECT POINT Gm = 2600
**X SELECT SWITCH SETTING = 6
**SIG SWITCH SETTING =  8
**METER FULL-SCALE Gm =  4000
**TUBE REJECT POINT METER INDICATION = 1950
**FACTORY-FRESH TUBE Gm % FULL-SCALE METER INDICATION = 100
**PROPORTIONAL uMHO MULTIPLIER =  1.333
```

The 100% full-scale indication for a factory-fresh 6AU6 is too high to accommodate any tolerances and variances that could result in an over-range meter indication. To accommodate this, the program allows the user to change the X Select and SIG switch settings for a more comfortable indication by means of the Action Menu following the above results:

```
*** ACTION MENU ***
 1) ENTER NEW X SELECT SETTING (0.5/1/2/3/4/5/6/7/8)
 2) ENTER NEW SIG SWITCH SETTING (1/2/4/8)
 3) DO ANOTHER RUN (different tube type)
 4) EXIT TO DOS
*ENTER <1-4> FROM ABOVE
```

In most cases, the best way to improve the full-scale meter indication is to change the X-Select value. To reduce the full-scale meter indication, we can choose the next lower X Select value (i.e., reducing it from the current value of 6 to 5). To do this, we enter selection <1> from the above action menu and respond as follows:

```
*ENTER NEW X SELECT SETTING (0.5/1/2/3/4/5/6/7/8)  5
```

The program then recalculates using the new input data with the following results:

```
**FACTORY-FRESH TUBE Gm = 4000
**65% RETMA TUBE REJECT POINT Gm = 2600
**X SELECT SWITCH SETTING = 5
**SIG SWITCH SETTING =  8
**METER FULL-SCALE Gm =  4800
**TUBE REJECT POINT METER INDICATION = 1625
**FACTORY-FRESH TUBE Gm % FULL-SCALE METER INDICATION = 83
**PROPORTIONAL uMHO MULTIPLIER = 1.600
```

Notice that the new full-scale meter indication is now 83% rather than 100%, which is a much more favorable result.

When changing the X Select or SIG switch settings, keep the following points in mind:

1. In most instances the best way to adjust the full-scale meter indication is to change the X Select (rather than SIG) switch setting. To reduce the full-scale meter indication, select the next lower X Select value (e.g., 5 rather than 6 as in the above example).

2. The "0.5" X Select switch setting is actually the CAL position (see Figure 23). This setting is required only for ultra-high Gm tubes.

3. In most cases it is unlikely that it will be necessary to change the SIG switch setting to obtain a more favorable full-scale meter indication. When running the initial computation, the program always selects the highest SIG switch setting possible (since this results in the lowest tube grid drive signal amplitude which in turn minimizes tube distortion).

4. For some high Gm tubes it will not be possible to find an X Select setting that results in a comfortable full-scale meter indication. As a case-in-point, for a tube with a Gm of 12,000 the best results that can be achieved by changing X Select switch are 50% or 100% full-scale meter indication. Even so, this is more of an inconvenience than a real problem since accurate Gm measurements can still be obtained using one or the other of these settings.

D. DETERMINING FACTORY-FRESH TUBE Gm

As per the above discussion, the computer program requires factory-fresh tube Gm as its primary input. This then opens the discussion as to how to best determine this number.

If you trust the roll-chart/tube list, factory-fresh tube Gm is easily determined by simply multiplying the listed R.P. value by the X multiplier and then dividing this product by 0.65. As a case in point, the roll-chart/tube list specifies an R.P. of 1300 and an X multiplier of 2 for a 6AU6. Factory-fresh Gm is then computed as (1300 x 2)/0.65 = 4000. This result is in good agreement with the 6AU6 Gm value (3900) published in the GE "Essential Characteristics" tube manual. (Note: For tubes where no X multiplier is specified, the implicit value is 1.)

E. ROLL-CHART/TUBE LIST VERSUS TUBE MANUAL

Although relying on the roll-chart/tube list is the simplest way to determine tube G_m (as well as the plate voltage and grid bias setups), I don't think it is very reliable. I've heard numerous complaints of "problem tubes" where users experienced off- or low-scale readings that they didn't believe.

Given this uncertainty, I believe that the most technically sound approach is to start from the ground up and determine G_m, plate/screen voltage, and grid bias voltage from a suitable tube manual. Fortunately, this is not a difficult task. A detailed explanation of this process is presented in Appendix F.

SECTION IX - CLOSING COMMENTS

At the outset of this project, my modest goal was to improve the TT-1A with incremental design changes to mitigate the issues that were causing significant transconductance measurement errors. As I gained a fuller understanding of how the TT-1A worked and the design issues responsible for these errors, I realized that a meaningful improvement would require a major instrument redesign. Even though this multiplied the project complexity by an order of magnitude, the additional required time and effort were well worth the results.

An additional benefit of this effort is that it forced me to gain a fuller understanding of the various issues associated with precision transconductance measurements (and a better understanding of vacuum tubes in general). I have made a best effort in this paper to pass these insights along to the readers.

Putting this into perspective, while the TT-1A is a capable mid-range vintage transconductance tube tester, the redesigned TT-1B is a fine precision instrument with exalted transconductance measurement accuracy *that can be verified by direct measurement with instrumentation available in most home electronics workshops*.

One of the reasons for the length and detail of this paper is that I believe this is all necessary for readers to be able to successfully replicate this redesign on their own TT-1As. Even so, I must caution that this project should be undertaken only by patient readers experienced in electronics assembly, testing, and troubleshooting.

Addressing my thoughts for a follow-on effort, my intent is to add a 4-digit numeric display with a 3-position selector switch to allow the following enhancements:

1. <u>Transconductance Measurement</u> - The numeric display will provide a more readable and precise transconductance indication. This will supplement (rather than replace) the existing TT-1A analog meter.

2. <u>Precise Filament Voltage Setting</u> - The numeric display will also serve as an AC voltmeter to allow more convenient filament voltage setting. (In the current implementation, this requires an external AC DVM.)

3. <u>Precise Grid Bias Voltage Setting</u> - Although the TT-1B as currently implemented improves grid bias voltage accuracy by the addition of a negative voltage regulator as per Section IV-B, the ultimate accuracy is still determined by the setability and linearity of the front-panel Bias control. To remedy this, the numeric display will also serve as a DC voltmeter to allow precise grid bias voltage setting.

This numeric display and selector switch will be mounted on the TTA-1-1 supplemental tube socket adaptor assembly (that mounts inside the lid of the TT-1A case).

As a final thought, it is likely a paper for a project this complex will contain some errors. I therefore invite and welcome any and all reader feedback covering possible errors or any other issues readers may wish to raise.

APPENDIX A - TT-1B_01.BAS COMPUTER PROGRAM EQUATIONS

This Appendix presents the various equations employed in the TT-1B_01.BAS computer program for the benefit of interested readers.

65% RETMA tube reject-point (R.P.) G_m:

$$RPG_m = 0.65 \times Gm$$

where:

RPG_m is the 65% RETMA tube reject-point value,
Gm is the factory-fresh tube transconductance.

Full-scale meter transconductance:

$$G_m = (3{,}000 \times SIG)/X$$

where:

Gm is the meter full-scale transconductance (in micromhos),
SIG is the Signal selector switch (signal oscillator) attenuator setting (1, 2, 4, or 8),
X is the X Select step attenuator setting (0.5, 1, 2, 3, 4, 5, 6, 7, or 8).

Tube reject-point meter indication:

$$TRP = G_m \times 0.65/(SIG/X)$$

where:

TRP is the tube reject point Proportional Micromho meter indication (on a scale of 0-3000),
Gm is the factory-fresh tube transconductance,
SIG is the Signal selector switch (signal oscillator) attenuator setting (1, 2, 4, or 8),
X is the X Select step attenuator setting (0.5, 1, 2, 3, 4, 5, 6, 7, or 8).

Factory-fresh tube GM as a percentage of full-scale meter reading:

$$\%GMFS = (G_m \times X)/(3000 \times SIG) \times 100$$

where:

%GMFS is the percent full-scale meter reading of a factory-fresh tube,
Gm is the factory-fresh tube transconductance,
X is the X Select step attenuator setting (0.5, 1, 2, 3, 4, 5, 6, 7, or 8),
SIG is the Signal selector switch (signal oscillator) attenuator setting (1, 2, 4, or 8).

Proportional Micromho multiplier:

$$PMult = SIG/X$$

where:

PMult is the multiplier factor required to convert the Proportional Micromho meter reading to actual Gm,
SIG is the Signal selector switch (signal oscillator) attenuator setting (1, 2, 4, or 8),
X is the X Select step attenuator setting (0.5, 1, 2, 3, 4, 5, 6, 7, or 8).

APPENDIX B - TT-1B ENHANCED CALIBRATION FEATURE

Readers familiar with the TT-1A calibration procedure can attest that it is a lengthy and tedious process that requires removing the unit from its case in order to access numerous internal adjustments. The TT-1B, in sharp contrast, includes an added front-panel CAL (calibration) feature than can be done on-the-fly in just a few seconds.

This new CAL feature effectively calibrates the TT-1B for a hypothetical tube having a transconductance of 24,000 micromhos (the maximum specified instrument transconductance measurement capability). For a tube with a transconductance of 24,000 micromhos (or 0.024 mhos for unit consistency) and the new TT-1B test load resistance (R_L) of 20.833 ohms, the voltage gain from equation (4) would be:

$$
\begin{aligned}
A_V &= G_m \times R_L \\
&= 0.024 \times 20.833 \\
&= 0.5
\end{aligned}
$$

With this in mind and referring to the functional block diagram of Figure 15, the signal voltage appearing across the 20.833 ohm R_L under these circumstances would be half that of the signal oscillator voltage applied to the tube grid. As per the TT-1B_01.BAS computer program and Appendix A, we would use an input signal level of 0.2 VPP (corresponding to a Signal selector switch setting of 8, the lowest drive level) and the lowest X multiplier (X=1) for a tube with such high transconductance.

Since the gain of a tube with a transconductance of 24,000 micromhos would be 0.5 into a 20.833 ohm R_L, a conceptually easy way to calibrate the instrument would be to insert a calibration signal of half the specified 0.2 VPP grid drive voltage, or 0.1 VPP, directly into the meter drive circuitry and then adjust the added front-panel CAL pot as required for a full-scale meter reading. As can be seen in the functional block diagram of Figure 15, this is facilitated by the OP/CAL switch.

The OP/CAL switch is actually a DPDT relay that is activated by the front-panel Line Test switch. As explained in Section V-F, the TT-1A Line Test function is no longer used or required for its original purpose in the TT-1B.

With this relay set to the CAL position, the calibration signal is directly connected to the signal amplifier (in place of the tube output). Since this signal is 0.1 VPP and the X multiplier (X Select) is 1 as per the preceding discussion, adjusting the new meter calibration pot (the CAL Adjust control as per Figure 15) for a full-scale meter deflection precisely calibrates the instrument.

For the sake of convenience, however, I decided to use the existing 0.2 VPP signal for calibration purposes (rather than adding a new 0.1 VPP calibration output) and then compensated for this by adding a special X multiplier selection (used for calibration purposes only) of 0.5. In effect then, this doubles the calibration drive signal amplitude and then cuts the meter drive circuitry gain in half to yield the same result as per Figure 15.

Since the 0.2 VPP calibration test signal is derived directly from the signal oscillator output attenuator voltage divider (and not through the Signal selector switch), the Signal selector switch can be left at any setting for the CAL function. To calibrate the TT-1B it is necessary only to set the new front-panel X Select switch to CAL (corresponding to X=0.5), pull down the Line Test switch, and then adjust the new front-panel CAL pot for a full-scale meter reading.

In addition to being easy to implement, this procedure is far more precise and yields much more stable results than was the case with the original TT-1A calibration procedure. Although the astute reader might point out that this new calibration feature only calibrates the instrument for a full-scale transconductance of 24,000 micromhos, the calibration is also accurate for all other ranges since 1) both the Signal (signal oscillator output) and X Select attenuators both use precision 1% resistors, and 2) the signal amplifiers and the meter itself are all linear.

Also note as per Figure 15 that the selectable 0.2/0.4/0.8/1.6 VPP signal oscillator output grid drive levels are nominal (approximate) values. Their exact values are established during the initial pre-calibration setup procedure where the signal oscillator output is adjusted slightly for a 20 VPP clean sinusoidal output at U1:B. This is done in the CAL mode with the X Select attenuator set for 0.5 (its CAL position). When this setup procedure is completed, the added front-panel CAL pot is then adjusted for a precise full-scale meter indication.

Readers should keep in mind that the TT-1B meter is most accurate when the instrument is placed in its specified horizontal position on a flat table or bench (as opposed to being set upright). Prior to calibration, the meter zero-set screw should be adjusted for a precise zero reading.

Aside from the X Select switch, other TT-1B switch settings do not affect calibration. However, *no tube should be plugged in to any of the test sockets during the calibration procedure*.

APPENDIX C - TT-1B TRANSCONDUCTANCE MEASUREMENT VERIFICATION

The point has been made on numerous occasions throughout this paper that there is no practical way to verify tube transconductance in the unmodified TT-1A based on objective signal voltage measurements. Essentially, the transconductance measurements of the unmodified instrument are “faith-based”, relying solely on confidence in the original design approach and not subject to easy verification. This serious shortcoming has been remedied in the TT-1B.

In order to measure tube transconductance by means of objective signal measurements that can be done with equipment available in a typical home electronics workshop, we need to know the following:

1. The tube input signal voltage (Vi).

2. The resulting tube output signal voltage (Vo).

3. The tube load resistance (R_L).

With this information, we can then compute the voltage gain Av easily enough by dividing the output signal voltage by the input signal voltage (Vo/Vi). Once done, we can then compute G_m using equation (3) as follows:

$Gm = Av/R_L$

Although we can objectively measure the input and output tube signal voltages in the predecessor TT-1A with an oscilloscope, R_L is very difficult to ascertain for the reasons discussed in detail elsewhere in this paper.

As also discussed in detail elsewhere in this paper, a major design improvement of the TT-1B is the inclusion of a constant and well-defined R_L (i.e., the new 20.833 ohm plate load resistor). In principle then, this solves the problem - we simply measure the input and output voltages as described above, compute the voltage gain Av, and then use equation (3) to compute G_m.

A practical complication is that since R_L is so low in value, very little output signal voltage is developed due to the resulting low gain. As an example, using a 6AU6 tube with a typical transconductance of 3800 micromhos, the circuit voltage gain computed using equation (2) would be only 0.079. Using the 6AU6 TT-1B specified grid drive level of 0.4 VPP, the signal output voltage developed across R_L would be only 0.032 VPP. This would be very difficult to measure without elaborate equipment, especially given the large amount of 50/60 Hz power supply ripple voltage present.

Referring to the functional block diagram of Figure 15, we can make an accurate proxy measurement of this signal voltage at the indicated G_m Verification Measurement Test Point. As per Figure 15, this test point is preceded by a high gain amplifier and two stages of highpass filtering to suppress the 50/60 Hz ripple voltage.

This measurement is most easily done with the unit in the CAL mode with an oscilloscope. Using the CAL mode in my TT-1B, I measured a signal voltage of 0.207 VPP (very close to the nominal 0.2 VPP calibration signal voltage) at the input of the first highpass filter (immediately following the OP/CAL relay) and 9.3 VPP at the Gm Verification Measurement Test Point (immediately following the differential amplifier output highpass filter) for a net gain of 9.3/0.207 = 44.9.

After performing the CAL procedure, I set up the TT-1B to test a 6AU6 tube, plugged the 6AU6 in, allowed sufficient time for a full warm-up, then measured its transconductance in the standard fashion with a resulting Proportional Micromhos meter indication of 1760. Applying the roll-chart X multiplier of 2 (which was verified by the TT-1B_01.BAS computer program to be correct), this corresponds to a true transconductance of 1760 x 2 = 3520 micromhos.

Again using the oscilloscope, I measured the 6AU6 grid drive voltage (this is conveniently done at the Signal attenuator switch SO pins 4 and 10) at 0.42 VPP (with the Signal switch set at 4 as per the roll-chart). Next, I measured the voltage at the Gm verification measurement test point at close to 1.40 VPP. The *apparent net voltage gain* (i.e., including the post-amplifier gain of 44.9) is thus:

$$
\begin{aligned}
Av &= Vo/Vi \\
&= 1.40/0.42 \\
&= 3.33
\end{aligned}
$$

Computing the apparent Gm (i.e., the Gm value including the post-amplifier gain of 44.9) using equation (3):

$$
\begin{aligned}
Gm &= Av/R_L \\
&= 3.33/20.833 \\
&= 0.159842 \text{ mhos} \\
&= 159{,}842 \text{ micromhos}
\end{aligned}
$$

Dividing this apparent Gm by the post-amplifier gain of 44.9, we obtain a true Gm of 159,842/44.9 = 3560 micromhos. Since this value is just slightly over 1% of that indicated on the TT-1B meter (3520), these results are in astonishingly good agreement.

The curious reader may wonder upon which of the two transconductance measurements should we should place the greatest reliance. Actually, this is really more of an academic issue since the results are so close (and well within the power of the human observation and estimation associated with the measurements). Even so, heaviest reliance should be placed up the measurement as indicated on the TT-1B meter (following the enhanced calibration procedure).

To explain, the TT-1B meter indication has the benefit of a carefully controlled calibration procedure that employs direct substitution of the signal source and 1% precision resistors. Essentially, this provides calibration laboratory precision.

Although the direct measurement technique yields good results, it requires more human judgment and interpretation. First, the oscilloscope signal amplitude reading requires some estimation. Second, the low-level grid drive signal is somewhat noisy, requiring yet additional estimation.

The important thing, however is that the two results are very close to each other, thus providing confidence via direct measurement that the design approach is valid.

Since this result seemed almost too good to be true, this same verification test was conducted on a 12AV6 (a triode rather than pentode requiring different setups). Once again, the transconductance measurement results were in excellent agreement (1450 vs. 1446 micromhos).

To facilitate this measurement for additional tubes, I installed two small coaxial DC connectors to the immediate right of the added CAL pot. (These connectors are not illustrated in the front-cover TT-1B photo.) One of these was wired to Signal attenuator switch (pin 4 for the signal, pin 10 for the common) to monitor the tube grid input signal voltage. The second was wired to the new signal processor G_m Verification Measurement Test Point (at the X Select switch; pin 10 for the signal and pin 1 for the common). Although these added connectors are used only for verification and are therefore optional, readers installing these connectors should select ones with plastic housings so that they are electrically insulated from the TT-1B front-panel.

With the benefit of having these connectors installed, I did confirmation measurements on a large number of tubes. In all cases these measurements were in good agreement with the TT-1B indicated transconductance.

APPENDIX D - INSTALLING THE TT-1B IN A PELICAN 1600 CASE

As originally supplied by Heathkit, the TT-1A was housed in a custom-designed wooden case. Although these wooden cases were satisfactory for their intended purpose of housing and protecting the TT-1A, they were prone to cosmetic damage with the result that many surviving TT-1As look worn-out and dilapidated.

Although the wooden case for my “primo” TT-1A was in good condition, my “junker” TT-1A wooden case was badly shop-worn. To remedy this, I decided to upgrade my “primo” TT-1A (now a TT-1B) wooden case to a modern and more rugged Pelican 1600 composite plastic case. I then placed my “junker” (reconditioned, but unmodified) TT-1A in the presentable Heathkit wooden case and discarded the dilapidated one.

The Pelican 1600 is a ruggedized, weather-proof case constructed of high-impact copolymer polypropylene. It is supplied with “pick and pluck” cubed foam inserts that allow users to conveniently create custom foam excavations. Although this case is larger than the original wooden case, it provides superior instrument protection and allows more room for accessories (as well as the follow-on 4-digit numeric transconductance display I intend to add).

Referring to the front-cover TT-1B photo , the TTA-1-1 supplemental tube socket adaptor assembly mounts inside the case lid (as it did in the original wooden case). A flexible pocket was installed (also inside the case lid) to contain the TT-1 operational manual and the most current tube list.

A large foam excavation was made to house the TT-1B, while a smaller one was made to house the AC power cord, the GE “Essential Characteristics” tube reference manual, and other miscellaneous items. Although not illustrated in the photo, a Velcro tie-down was installed inside the case lid to secure the two TTA-1-1 extension cables.

APPENDIX E - TT-1B_01.BAS SOURCE CODE LISTING

The following TT-1B_01.BAS source code listing must be run under GW-BASIC, QBASIC, QuickBASIC or other compatible versions of BASIC. The compiled version (TT-1B_01.EXE) runs directly as an executable file. Both of these files can be downloaded from the N6DC vintage radio website.

This software, as originally compiled, does not run on current Windows PC operating systems and must be run on a legacy PC that supports MS-DOS software. In addition to MS-DOS computers, it can also be run under Windows 3.1, Windows 95/98, and Windows ME. When running this software under Windows 95/98, use the MS-DOS prompt (window) or true MS-DOS mode.

Since using a legacy PC is likely to be inconvenient for most readers, a new version of this software has been compiled (TT-1B_01W.EXE) that is directly executable under current Windows operating systems. To run this Windows version, just click on this file name and the program window will open. Once opened, program operation is identical to that of the MS-DOS version. TT-1B_01W.EXE can be downloaded from the N6DC vintage radio website.

The compiler used to convert TT-1B_01.BAS to a Windows-executable file is "QB64". This is a freeware utility written to compile legacy MS-DOS BASIC programs to Windows-executable formats. This compiler can be downloaded from http://www.qb64.net.

The TT-1B_01.BAS source code listing is as follows:

```
1000 REM ***** TT-1B_01.BAS (11-02-2013) *****
1010 REM ***** by Alex J. Burwasser, N6DC *****
1020 REM
1030 REM
1040 REM
1050 REM ***** VARIABLES *****
1060 REM SIG - TT-1B signal switch setting
1070 REM X - TT-1B X Select switch setting
1080 REM XX - X value rounded-down to nearest integer (1-8; 0.5 for 0)
1090 REM FSGM - Meter full-scale Gm (micromhos)
1100 REM TRP - Tube Reject Point meter indication (65% of factory-fresh tube Gm in
micromhos)
1110 REM PCFSGM - % full-scale meter reading for factory-fresh tube
1120 REM PMULT - Proportional Micromho multiplier (multiplies Proportional Micromho
1130 REM         meter reading to actual Gm)
1140 REM RPGM - Tube 65% RETMA Reject Point Gm
1150 REM
1160 REM
1170 REM
2000 REM ***** OPENING SCREEN *****
2010 COLOR 15, 1, 1: CLS
2020 PRINT "*** TT-1B_01.BAS ***                    By Alex J. Burwasser, N6DC
(11-02-2013)"
2030 PRINT
```

```
2040 PRINT "This program allows the user to compute proper TT-1B X Select and SIG
switch      settings for most tubes.  The user is first prompted to enter the tube
factory-    fresh Gm value.  The program then computes recommended X Select and SIG
switch
2050 PRINT "settings.  It also computes the tube RETMA 65% reject point (R.P.) Gm,
meter       full-scale Gm, tube reject point meter indication, meter full-scale Gm,
factory-    fresh tube Gm as a percentage of meter full-scale, and the Proportional
Micro-
2060 PRINT "mho multiplier that converts the meter reading to actual Gm.  An action
menu is     then presented that allows the user to specify different X Select or SIG
switch      values if desired.
2070 PRINT
2080 INPUT"*PRESS <RET> TO CONTINUE OR ENTER <X> TO EXIT TO DOS  ",P$:PRINT
2090 IF P$="X" OR P$="x" THEN SYSTEM
2100 REM
2110 REM
2120 REM
3000 REM ***** DATA ENTRY *****
3010 INPUT"*ENTER FACTORY-FRESH TUBE Gm (375-48,000 uMHOs)  ",GM
3020 IF GM<375 OR GM>48000! THEN BEEP:GOTO 3010
3030 PRINT
3040 GOSUB 5000
3050 GOSUB 6000
3060 PRINT
3070 REM
3080 REM
3090 REM
4000 REM ***** ACTION MENU *****
4010 PRINT"*** ACTION MENU ***"
4020 PRINT" 1) ENTER NEW X SELECT SETTING (0.5/1/2/3/4/5/6/7/8)
4030 PRINT" 2) ENTER NEW SIG SWITCH SETTING (1/2/4/8)
4040 PRINT" 3) DO ANOTHER RUN (different tube type)
4050 PRINT" 4) EXIT TO DOS"
4060 INPUT"*ENTER <1-4> FROM ABOVE  ",P$
4070 IF P$="1" THEN PRINT:GOSUB 7000:GOSUB 6000:PRINT:GOTO 4000
4080 IF P$="2" THEN PRINT:GOSUB 7000:GOSUB 6000:PRINT:GOTO 4000
4090 IF P$="3" THEN CLS:GOTO 3000
4100 IF P$="4" THEN SYSTEM
4110 BEEP:GOTO 4060
4120 REM
4130 REM
4140 REM
5000 REM ***** COMPUTATIONS GOSUB (for Gm entry) *****
5010 SIG=8
5020 X=3000*SIG/GM
5030 XX=INT(X)
5040 IF X>8 THEN SIG=SIG/2:GOTO 5020
5050 IF XX=0 THEN XX=.5
5060 FSGM=3000*SIG/XX
5070 TRP=GM*.65/(SIG/XX)
5080 PCFSGM=GM/FSGM*100
5090 PMULT=SIG/XX
5100 RPGM=GM*.65
5110 RETURN
5120 REM
5130 REM
5140 REM
6000 REM **** DISPLAY RESULTS ON SCREEN *****
6010 PRINT"**FACTORY-FRESH TUBE Gm = ";GM;"
```

```
6020 PRINT"**65% RETMA TUBE REJECT POINT Gm = ";INT(RPGM+.5)
6030 PRINT"**X SELECT SWITCH SETTING = ";XX
6040 PRINT"**SIG SWITCH SETTING = ";SIG
6050 PRINT"**METER FULL-SCALE Gm = ";INT(FSGM+.5)
6060 PRINT"**TUBE REJECT POINT METER INDICATION = ";INT(TRP+.5)
6070 PRINT"**FACTORY-FRESH TUBE Gm % FULL-SCALE METER INDICATION = ";INT(PCFSGM+.5)
6080 PRINT"**PROPORTIONAL uMHO MULTIPLIER = ";
6090 PRINT USING"##.###";PMULT
6100 RETURN
6110 REM
6120 REM
6130 REM
7000 REM ***COMPUTATIONS GOSUB (for X Select or SIG input)
7010 IF P$="1" THEN INPUT"*ENTER NEW X SELECT SETTING (0.5/1/2/3/4/5/6/7/8) ", XX
7020 IF XX<>.5 AND XX<>1 AND XX<>2 AND XX<>3 AND XX<>4 AND XX<>5 AND XX<>6 AND XX<>7
AND XX<>8 THEN BEEP:GOTO 7010
7030 IF P$="2" THEN INPUT"*ENTER NEW SIG SWITCH SETTING (1/2/4/8) ", SIG
7040 IF SIG<>1 AND SIG<>2 AND SIG<>4 AND SIG<>8 THEN BEEP:GOTO 7030
7050 FSGM=3000*SIG/XX
7060 TRP=GM*.65/(SIG/XX)
7070 PCFSGM=GM/FSGM*100
7080 PMULT=SIG/XX
7090 RPGM=GM*.65
7100 PRINT
7110 RETURN
7120 REM
7130 REM
7140 REM
7150 END
```

APPENDIX F - DETERMINING FACTORY-FRESH TUBE G_m & ELECTRODE VOLTAGES

Determining factory-fresh tube G_m is necessary since this is the primary input to the TT-1B_01.BAS computer program. To be completely thorough, first do the following preliminary steps:

1. Confirm the Correct Tube Filament Voltage - This information can be found in an appropriate tube manual or data sheet. (Most tube data sheets can now be found on-line.) Also, the filament voltages specified in the Nov 1978 final edition tube list (available in PDF format from the N6DC vintage radio website) are likely all correct.

2. Confirm the Tube Socket Base Diagram - This is very important since it establishes the nine TT-1B Selector switch settings (keeping in mind that incorrect settings can damage a tube). The base diagram can be found in a tube manual or data sheet. Also, the tube socket base diagrams specified in the Nov 1978 final edition tube list are probably all likely all correct.

Once above steps are done, it is necessary to determine the plate/screen and grid bias voltages:

1. TT-1A Plate Voltage Selector Switch Setting - The Plate switch selects the following plate supply voltages for the tube under test:

   0 - 26 VDC
   A - 20 VAC *
   B - 45 VAC *
   C - 90 VDC
   D - 135 VDC
   E - 177 VAC *
   F - 225 VDC
   G - 80-220 VDC **

   * AC voltages are used for diodes, rectifiers, and some special-function tubes.

   ** Variable DC voltage is used for voltage regulator tubes.

   For triodes, determining an appropriate Plate switch setting is straightforward. To illustrate this by example, I looked up the 6C4 in the GE "Essential Characteristics" tube manual and found that it is rated for Class A amplifier service at plate voltages of both 100 (for 0 volts grid bias) and 250 VDC (for -8.5 VDC grid bias). Although voltages close to both of these values are available from the Plate voltage selector switch (positions C and F, respectively), I chose the higher voltage (position F) to avoid the uncertainties and complications of running the tube at 0 volts grid bias. (Position F is also the setting specified in Nov 1978 final tube list.)

   At this higher voltage, 6C4 factory-fresh tube G_m is specified as 2200 micromhos. This number is in perfect agreement with that computed directly from the tube list as per Section VIII-D.

   For tetrodes pentodes, and other multi-grid tubes, there is an additional complication.

Although the TT-1A applies the selected plate voltage to both the plate and screen of the tube under test, tube manuals often specify different plate and screen voltages.

To illustrate this by example and beginning first with the “easy” scenario where a tube manual listing can be found with the *same* specified plate/screen voltages, I looked up the 6AU6 (once again in the GE “Essential Characteristics” tube manual) and found that it is rated for Class A amplifier service with the plate and screen voltages both set at 100 VDC. This is a good match for the position C Plate voltage selector switch setting of 90 VDC (and also what is specified in the Nov 1978 final tube list).

At 100 VDC, 6AU6 factory-fresh tube G_m is specified in the tube manual as 3900 micromhos. This number is in good agreement with that computed directly from the tube list (4000) as per Section VIII-D.

The more difficult scenario is that where the tube manual specifies *different* plate and screen voltages. A good case in point is the 6AG7. Again referring to the GE “Essential Characteristics” tube manual, the 6AG7 is rated for Class A amplifier service for plate and screen voltages of 300 and 150 VDC, respectively. As per the above discussion, however, the TT-1A is designed to supply the same voltage to both the plate and screen grid. (Although we can select this voltage using the Plate switch, the TT-1A does not provide *separate* plate and screen voltages.)

My working assumption is that the screen voltage has the predominant effect on G_m. To informally confirm this, I searched all 6.3 volt filament tubes in the GE “Essential Characteristics” tube manual to find multi-grid tubes specified with two or more plate voltages *at the same screen voltage* (with the provisos that the plate voltage equaled or exceeded the screen voltage and that the negative grid biases were similar). In all cases, the specified G_m values were very close. Specific tubes found in this search included the 6AK5, 6BA6, 6BE6, 6BJ6, 6J7,6K6, 6KV8, 6SA7, and 6SJ7.

This being the case, *we set the Plate voltage selector switch to whatever plate/screen voltage best matches the tube manual specified* ***screen*** *voltage.* The setting that bests matches the listed 6AG7 150 VDC screen voltage specification is thus D (135 VDC). This is also the setting specified in the Nov 1978 final tube list.

Under these conditions, 6AG7 factory-fresh tube G_m is specified in the tube manual as 11,000 micromhos. This number is in excellent agreement with that computed directly from the tube list (11,262) as per Section VIII-D.

To succinctly summarize this procedure, always select a TT-1A Plate voltage selector setting that best matches the typical operating voltages specified in the tube manual or data sheet. For multi-grid tubes, try to find a tube manual listing where the typical operating voltages are the same for both the plate and screen. If this is not possible, then select a TT-1A Plate voltage selector setting that best matches the specified screen voltage. Also, avoid plate voltage settings called for in the tube manual where zero grid bias is specified.

Although not yet addressed, this procedure requires that the grid bias voltage also be appropriately set. This issue is discussed in the following paragraph.

2. <u>Determine the Grid Bias Voltage</u>

   In addition to specifying appropriate plate and screen voltages, the tube manual also specifies a corresponding negative grid bias voltage. This is often specified directly, as is the case for the -8.5 VDC grid bias specified for the 6C4 triode from the above example.

   In other cases, this voltage is specified indirectly as a cathode resistor value. Referring again to the 6AU6, the tube manual does not list the grid bias voltage but instead specifies a cathode resistor of 150 ohms for the 6AU6 test setup discussed above.

   Fortunately, only simple arithmetic is required to compute the corresponding negative grid bias. Specifically, we multiply the specified cathode resistor value by the total *cathode* current. Keeping in mind that total cathode current for a screen grid tube is the plate *plus* the screen currents, this comes to 5.0 mA plate current plus 2.1 mA screen current (as per the tube manual) for a total cathode current of 7.1 mA. Multiplying this cathode current by the specified cathode resistor value, we obtain 0.0071 amperes x 150 ohms = 1.065 volts. This is close to the 0.9 volt (negative) value specified in the Nov 1978 final tube list for the 6AU6.

   For triodes, of course, this slight subtlety does not have to be addressed. As long as no control grid current is drawn, the cathode current is the same as the plate current.

   Some additional arithmetic is required to convert the listed or calculated negative grid bias voltage to the appropriate TT-1A Bias control setting.

   As per the Heathkit TT-1 operational manual, the Bias control is a linear potentiometer with a calibrated scale of 0-50. In addition, there is an adjacent H/L (high/low) bias Range selector switch. When this Range switch is set to H (high), the Bias control output is 0 to -20 VDC. For the L (low) range switch setting, the Bias control output is 0 to -5 VDC.

   For the High bias range then, a Bias control setting of 10 (for example) corresponds to an applied grid bias of -4.0 VDC. For the Low bias range, a Bias control setting of 10 corresponds to an applied grid voltage of -1.0 VDC.

Once the factory-fresh tube G_m has been determined and the appropriate plate voltage selector switch setting and grid bias value has been established, run the TT-1B_01.BAS computer program as discussed above to determine the correct X Select/SIG switch settings, R.P. value, and Proportional Micromho multiplier. With this information in hand, the tube is ready to be tested. This information can then be annotated on the appropriate tube line in the tube list for future reference.

As a post-note to this section, I believe that in many cases the TT-1 understates G_m as a result of the selected plate/screen voltage being somewhat lower than the more typical values specified in the tube manuals. More specifically, in cases where the TT-1 specified plate/screen voltages are 90, 135, and 225 VDC, the tube manual specified voltages are frequently 100, 150, and 250 VDC, respectively. In general, higher plate voltages result in higher G_m. On a related note, an additional possible improvement to the TT-1B would be to adjust the plate/screen voltage to values matching those in the tube manual (i.e., 100, 150, and 250 VDC rather than 90, 135, and 225 VDC, respectively).

APPENDIX G - UPCOMING ENHANCEMENTS FOR THE TT-1B

Steve Mazur and I have already embarked on a project to enhance and modernize the TT-1B that will add new features, further increase precision, and simplify operation. To briefly summarize the enhancements under consideration:

1. **Digital Metering (electrode voltages)** - The TT-1B in its current implementation does not have the means to conveniently meter the grid bias, plate, and filament voltages. We plan to remedy this using independent LED digital panel meters so that these voltages can be monitored conveniently, accurately, and simultaneously. (Simultaneous monitoring of the plate and filament voltages is especially desirable since their settings are interactive.)

Figure 24 - TT-1B Upgrade to Numeric Displays

2. **Digital Metering(Gm)** - Although the TT-1B indicates tube Gm (actually “Proportional Micromhos” to be completely accurate) by means of its large analog panel meter, the enhanced TT-1B will include an LED digital panel meter for better presentation and accuracy. The existing analog panel meter will still remain functional. Since the roll-chart serves no purpose in the TT-1B, it will be removed so that the added digital panel meters can be installed in the roll-chart panel cut-out.

3. **Variable Plate Voltage Capability** - The TT-1B in its current implementation has the same four DC plate voltage selections as the TT-1/TT-1A (26, 90, 135, and 225 VDC). These voltages, however, are only approximate and are unregulated. The relevant issue is that tube Gm varies with all of its electrode voltages (plate, bias, and filament). Although the enhanced TT-1B will allow accurate digital panel metering of all of these voltages and retain the means to set the bias and filament voltages, it will also include the added means to adjust the plate voltage as well.

   To elaborate, tube data sheets specify Gm for specific values of electrode voltages. To measure Gm accurately based on these data sheets, the tube tester must have the means to closely match all of these specified electrode voltages. By including the means to adjust the plate voltage, the enhanced TT-1B will have the flexibility to accurately set all specified tube electrode voltages.

4. **Audio Output Jack** - Sometimes tubes that otherwise test satisfactorily are “noisy”, or exhibit excessive microphonics when tapped or jarred. An audio output jack is useful in testing for such noisy tubes.

APPENDIX H - MODERNIZED WINDOWS USER SOFTWARE

A modernized and greatly improved user software program has been written to facilitate easier and more convenient Gm computation. The use of this new software is demonstrated by the test procedure below for a 6AU6A sharp-cut-off pentode:

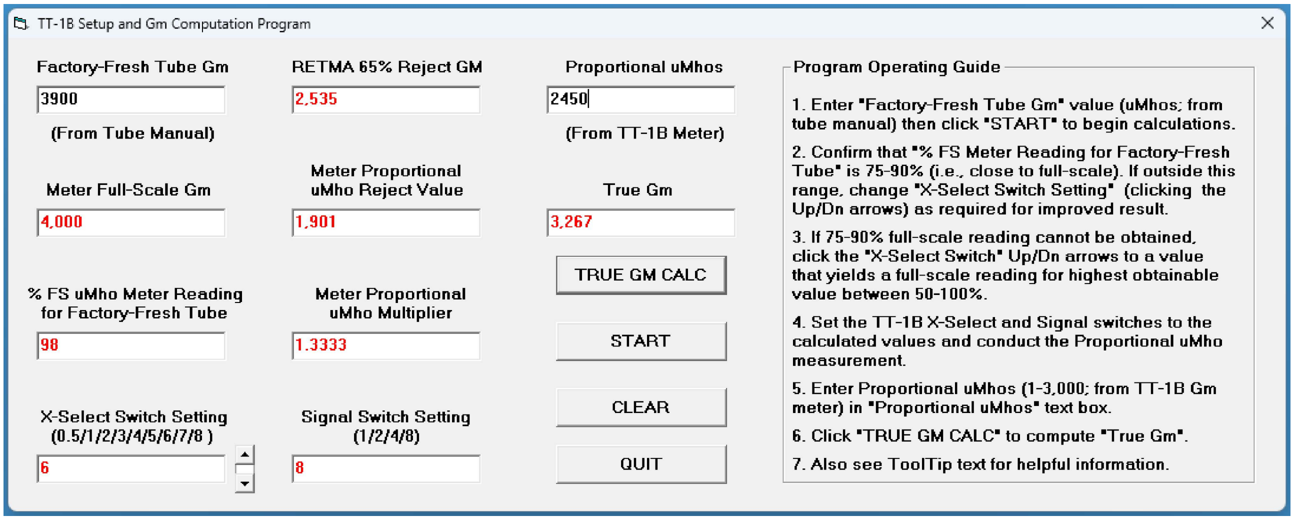


Figure 25 - TT-1B Setup and Gm Computation Program Main Screen

1. Refer to the GE Essential Characteristics (or other suitable) tube manual (see reference 6) to determine suitable plate, screen, grid bias, and filament voltages for a 6AU6A. Using the GE Essential Characteristics data, the 6AU6A can be operated as a class A amplifier with both the plate and screen voltages at +100 VDC, a grid bias of -1.065 VDC, and a filament voltage of 6.3 VAC. With these voltages, the factory-fresh tube Gm is specified as 3,900 micromhos.

2. Enter 3900 as the Factory-Fresh Tube Gm text box as illustrated on the program main screen in Figure 25 above.

3. Click “START” to begin the setup calculations. The results are illustrated by the text boxes displaying the red numerals.

4. Set up the TT-1B for a 6AU6A tube measurement using the plate, screen, grid bias, and filament voltages from the tube manual as discussed above. *Note that the only roll-chart/tube list settings used are the “Selectors” switch settings, plus the K1, P1, and G1 settings and plate/grid caps (if used).*

5. Since the program specifies a TT-1B X-Select switch setting of 6 and a Signal switch setting of 8 as per Figure 25, set these TT-1B switches accordingly.

6. Confirm that all TT-1B settings (including the Selectors) are correct.

7. Conduct the TT-1B Gm measurement. For the 6AU6 used for this test, the indicated Proportional MicroMhos (as read from the 0-3,000 TT-1B meter) was 2450.

8. Finally, enter the Proportional MicroMhos (2450 for the 6AU6A used for this test) in the program Proportional MicroMho text box as illustrated above and click the TRUE GM CALC button. This causes the program to convert the 2450 Proportional MicroMho value into true Gm. This yields a true Gm of 3,267, which is displayed in red numerals in the True Gm text box.

Using the RETMA 65% reject Gm as the pass/fail criterion (2,535 as indicated in the RETMA 65% Reject Gm text box), this tube passes.

The remaining text box numbers are for reference only. Also refer to the Program Operating Guide (on the screen right side) and the tooltip text for additional helpful information.

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