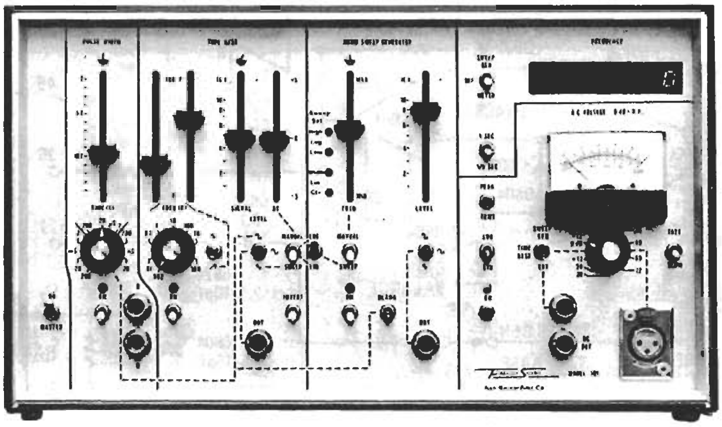
Audio Test Station



Part 4—This month we discuss in detail the theory of operation and the construction of the voltmeter and digital frequency meter that are essential to maximum usage of this audio test station.

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THIS MONTH, WE CONCLUDE THE CONSTRUCTION details of the Audio Test Station with the voltmeter and frequency counter sections.

Voltmeter section

The voltmeter circuit shown in Fig. 10 has two separate inputs. One is labelled EXT INPUT, which is a high-impedance low-gain input for line-level signals and the other is labelled MIKE INPUT, which is a low-impedance, high-gain input for low-level low-impedance signals. The front panel connector for the MIKE INPUT (J6) is a standard XLR type. The mike input circuitry consists of IC601 and its associated components, which form a balanced-line receiver. Resistor R601 terminates the line and R603 and R602 set the gain of IC601. The gain as seen from pin 3 of J6 is 16. The gain of the + input of IC601 is 16 + 1 or 17. Therefore, R603 and R604 are added to form a voltage divider so the overall gain of a signal applied at J6 pin 2 is the same as for a signal applied at J6 pin 3. This satisfies the basic purpose for having a balanced transmission line, namely common-mode rejection.

The balanced transmission line reduces the effect of stray signals picked up by interconnecting cables. The most susceptible stray pickup point in sound systems is a microphone cable. It is not that the mike cable itself is more susceptible, but rather that the microphone supplies such a low-level signal that it is relatively easy to pick up enough of a stray signal to be audible when mixed and amplified with the mike signal. The stray signal is often induced into a mike cable when the cable is in the presence of an electric or magnetic field.

The mike input circuitry inverts the signal applied to J6 pin 3, while the signal applied to pin 2 is not inverted. The actual mike signal that is applied to J6 pin 3 is of equal magnitude and opposite polarity of that which is applied to pin 2. Op-amp IC601 inverts the negative signal at pin 3, and adds this to the positive signal at pin 2.

When stray signals that are induced into the cable arrive at J6 pins 2 and 3 in phase, the signal at pin 3 is inverted and added to that at pin 2. However, this is subtraction. The result is what is known as common-mode rejection. A signal that was common to both inputs is rejected.

This circuit is normally shown with the four resistors of high precision and fixed value. In this circuit the common-mode rejection is simply trimmed by R604. To do this, connect a signal simultaneously at J6 pins 2 and 3. A mid-frequency sinewave at about a half a volt peak-to-peak is suitable. Set S17 to −72 dB, adjust R604 for minimum output from IC603. The signal at the output of IC603 should be about the same magnitude as the input signal. Note that the same signal applied to either input alone would have been amplified by about 4,000 or 72 dB. Therefore, since the signal was passed with a gain of approximately 1, the common-mode rejection ratio is the 72 dB that the signal would ordinarily have been amplified by.

We have used this type of mike preamp in sound systems with mike cable lengths of well over 100 feet without any apparent extraneous pickup. However, for sound system use, the 748 is much too noisy. This can be easily remedied though by replacing the 748 with something like TI's TL071. The two parts are pin-compatible, and the 5-PE compensation capacitors used for the 748 can even be left in place when switching to the TL071. The difference in noise levels is dramatic.

A second-stage preamp consisting of IC602 is included. Having the preamp in two stages accomplishes a couple of things. First, the supply voltages are rather close to the maximum output voltage. A single amplifier with 48-dB gain and with an attenuator at the output would clip at signal amplitudes intended for the low-gain attenuator setting. In fact a signal which yields full scale on the −30-dB setting causes IC602 to clip.

Another reason for cascading is that the overall bandwidth is increased. This allows two common op-amps to be used in place of a single high-frequency unit.

Switch S16 selects one of three high-level inputs. The sweep generator and timebase pickup points are on the inboard side of the fuse that is used in conjunction with the output overvoltage sensors. Therefore, if there is no output from the sweep generator or timebase but its level can be read on the meter, it indicates a blown fuse. Capacitor C603 blocks any DC component present at S16. Resistors R610 through R615 form the actual input attenuator. Consider for the moment just this resistor string and S17. There is stray capacitance between every node of that resistor string and every other node. The result is a very complex R–C filter. At low frequencies, it is a simple voltage divider whose transfer function can be easily determined by simple arithmetic. Higher frequencies, however, travel from node to node in a manner that is quite difficult to predict accurately.

Capacitors C604 through C609 are added as an attempt to swamp out the internodal stray capacitance. This is

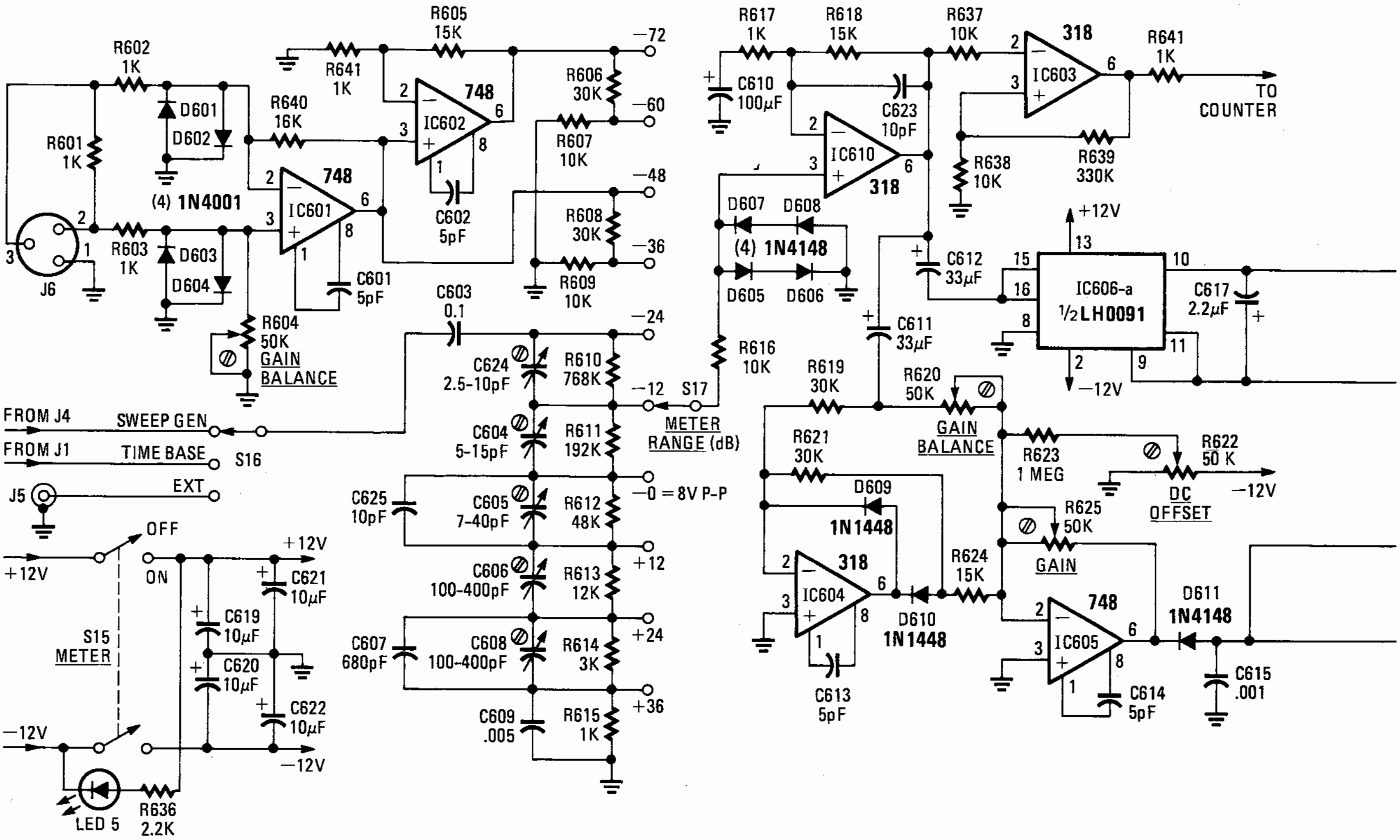


FIG. 10—SCHEMATIC DIAGRAM of the voltmeter. Its input circuitry uses frequency compensation. The separate preamp is for low-level inputs. The 1.5K resistor near the meter is R643, not R635.

called attenuator compensation. For purposes of simple compensation, the absolute capacitance values are relatively unimportant except that they must properly relate proportionately to the resistors that they are parallel with and the stray capacitances they are combining with. However, it is generally desirable to keep these capacitors small so as to minimize the total capacitance seen from the input.

If this were the input attenuator to an oscilloscope where waveform purity is of great importance, a low capacitance switch would be necessary. This is typically a large open-frame type with relatively large intercontact spacing. In this system, the problem is somewhat intensified by the use of a miniature switch where all nodes are in very close proximity. However, it is quite satisfactory up to at least 100 kHz. which should satisfy the needs of the system.

To adjust the compensating capacitors, monitor the output of IC603 with a scope. Apply a 10–15-kHz squarewave to the input and move S17 through each of the relevant positions. Adjust the capacitor immediately below the selected switch position for best squarewave response. Initial calibration of the attenuator will require several trials to bring the trimmers into proper relationship with each other. Once the adjustment is made with a squarewave input, a sinewave can be applied and swept from 10 kHz to 100 kHz to verify the overall response. Some fine tuning may be helpful at this time.

Assuming the sinewave from the timebase would be used for this last check, recall that the timebase begins a noticeable rolloff around 100 kHz. So be sure that any compensation adjustments made with the sinewave input are not compensating for timebase rolloff.

An additional gain stage and attenuator buffer consisting of IC603 is provided. Diodes D605 through D608 provide protection against excessive voltage applied to the external input. With S17 in the −24-dB position, IC603 is connected essentially straight to the external input. The diodes conduct at about 1 volt and are capable of 50 milliamps. Resistor R616 is intended to limit the diode current to a safe level. However, R616 cannot be arbitrarily large since it combines with the capacitance at the input of IC603 to form a low-pass filter. Since the diodes can handle 50 milliamps, 500V can be applied without damaging them. However, the power-handling capability of R616 must be considered at a fraction of that voltage level. If R616 is a half-watt resistor, it will tolerate 70 volts. The resistor can tolerate larger voltage levels than that for short periods of time. Since even a much smaller level will cause the meter to peg, if the operator is attentive, a 10K half-watt resistor should be adequate.

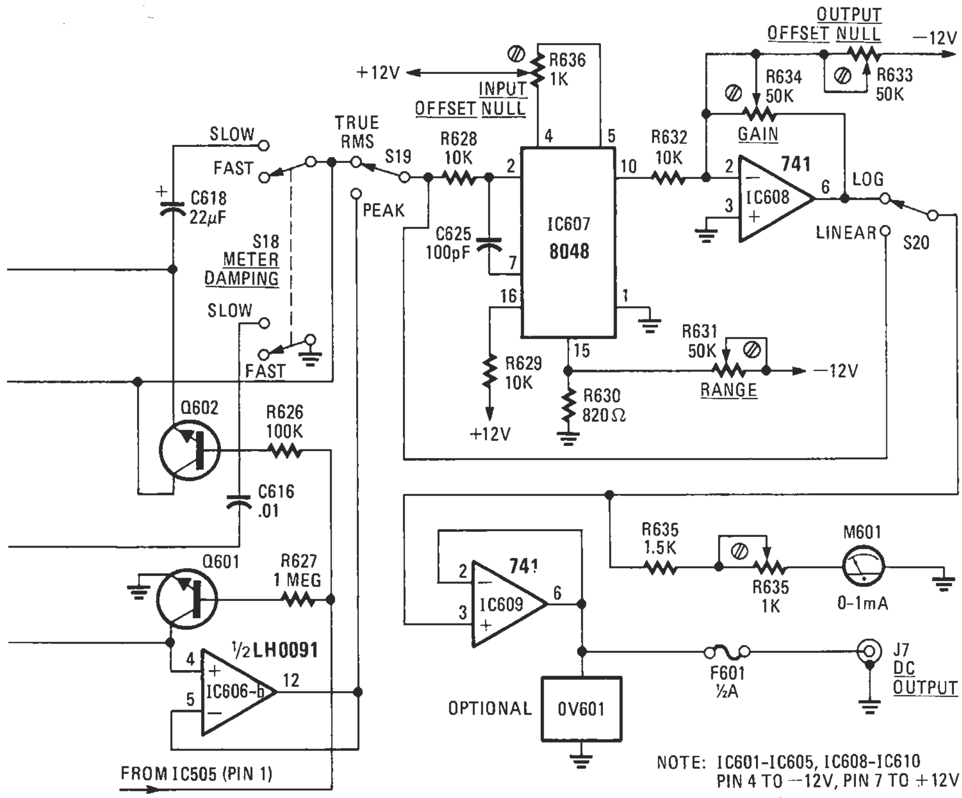
However, for insured safety R16 can be replaced with a positive temperature coefficient thermistor of approximately 10,000 ohms cold resistance. Such a device is a self-protecting resistor, since as excessive power causes it to heat, its resistance increases, which decreases the power and hence the heating effect. Also, keep in mind that R616 is of no particular value and need only be large enough to protect the diodes and small enough not to degrade high-frequency performance.

IC610 is a comparator with hysteresis. A zero-crossing dejector is formed by applying a signal through an isolation resistor to the negative input. Applying positive feedback causes the detection point to shift, which results in a circuit that ignores signals smaller than the detection levels.

This circuit preconditions the voltmeter signal before it is applied to the counter. For the circuit shown, the overall result is that the counter will count an input whose magnitude is 10% of the full-scale setting selected by S17.

IC604 through IC607 form the AC-to-DC converter. IC604 and IC605 are a full-wave rectifier. One-half of an applied waveform is inverted and added to the noninverted half. Resistor R630 adjusts the relative gain of the inverted and noninverted parts of this waveform. Trimmer R625 adjusts the overall gain of the rectified waveform.

Components D611, C615 and IC606 form a peak detector. When the signal from IC605 goes positive, diode D611 conducts charging C615. Capacitor C615 is small and hence charges as fast as the voltage from IC605 rises. When the output of IC605 passes a peak and starts to decrease, diode D611 cuts off and IC606 is left with charged C615 on its input.



Capacitor C615 discharges through the input of IC606. The rate at which C615 discharges is called the droop rate. Note that when D611 is cut off, IC606 has no DC bias on the + input. This results in a negative offset from IC606 equal to 1 diode-drop. Trimmer R622 shifts the output of IC605 enough to compensate for this and hence zero the output of IC606.

Note that the three diodes in the circuit do not exhibit the 7/10-volt threshold normally associated with a diode. The threshold is reduced to a negligible level by the active circuitry with which the diodes are associated.

The result so far then is a full-wave peak detector that will track an increasing amplitude instantaneously and track a decreasing amplitude at a slower rate, this rate being determined by the capacitor, and hence the droop rate is rather arbitrary and is based on the system under test.

If the acoustics of a large reverberant room are being analyzed, standing waves can cause wide fluctuation in amplitude. Tracing these rapid fluctuations requires a fast droop rate and hence a small capacitor. On the other hand, a fast droop rate will tend to track individual waveforms of a low-frequency signal. Therefore, to provide for difference response time requirements, S18 is provided to switch in a parallel capacitor, C616.

Also recall from the generator sections that there is a blanking mode that switches off the output of the audio generator at the end of a sweep. The same trigger that switches off the output of the audio generator turns on Q601, which rapidly discharges the capacitor at the input of IC606. Therefore, a clean cutoff is provided at the end of a plot.

The same signal that is applied to the peak detector is also applied to IC607. This is a true RMS-to-DC converter. Capacitor C617 controls the amount of filtering of the DC output. Unlike the peak detector, the response of IC607 is equal regarding both increasing and decreasing signal levels. Again S18 provides for a variable response time by adding C618 in parallel. Also the blanking trigger turns on Q602 to rapidly zero the output at the end of a sweep.

IC607 is a linear-to-log converter. Voltage gain in dB is defined as 20 times the log of the linear voltage gain. There is a linear voltage at S19 that is zero for a zero level input and ranges up to 4 volts for either a 4-volt peak or RMS input. Since a true log function cannot be physically implemented because it is of infinite length, a decision must be made to use only part of the function and eliminate the region that extends below some practical limit.

To help select an appropriate range; note the effect of a log converter on a linear input signal. Input variations slightly above zero produce very large output variations. Therefore, such things as ripple from the peak detector would be greatly magnified at low levels. Eliminating the lower portion of the function produces a dead zone, for any input voltage below the cutoff point is ignored. The cutoff-point trade off is between low-level excessive sensitivity and total dynamic range, which is the remaining portion of the function. Thus, it was rather arbitrarily decided to cut off the function at minus 40 dB.

VOLTMETER PARTS LIST

Resistors, ¼ watt, 5% or better
R601, R603, R617, R641—1000 ohms
R604, R620, R622, R625, R631, R633, R634, R640—50,000 ohms, trimmer
R605, R624—15,000 ohms
R606–R608—30,000 ohms, ¼ watt, 1%
R607, R609—10,000 ohms, ¼ watt, 1%
R610—768,000 ohms, ¼ watt, 1%
R611—192,000 ohms, ¼ watt, 1%
R612—48,000 ohms, ¼ watt, 1%
R613—12,000 ohms, ¼ watt, 1%
R614—3000 ohms, ¼ watt, 1%
R615—1000 ohms, ¼ watt, 1%
R616, R628, R629, R632, R637, R638—10,000 ohms, ½ watt
R618, R621—30,000 ohms
R623, R627—1 megohm
R616—100,000 ohms
R630—820 ohms
R635, R636—1000 ohms, trimmer
R639—330,000 ohms
R642—2200 ohms
R643—1500 ohms

Capacitors
C601, C602, C613, C614—5 pF disc ceramic
C603—0.1 μF disc ceramic
C604—5—15 pF trimmer
C605—7—40 pF trimmer
C606, C608—100—400 pF trimmer
C607—680 pF disc ceramic
C609—.005 μF disc ceramic
C610—100 μF electrolytic, 16 volts or higher
C611, C612—33 μF, electrolytic, 16 volts or higher
C615—.001 μF disc ceramic
C616—0.01 μf disc ceramic
C617—2.2 μF electrolytic or tantalum
C618—22 μF electrolytic, 16 volts or higher
C619-C622—10 μf electrolytic, 16 volts or higher
C623—10 pF disc ceramic
C624—2.5—10 pF disc ceramic
C625—100 pF disc ceramic
C626—10 pF disc ceramic

Semiconductors
IC601, IC602, IC605—748
IC603, IC604, IC610—318
IC606—LH0091 (National)
IC607—8048 (Intersil)
IC608, IC609—741
D601-D604—1N4001
D609-D611—1N4148
Q601, Q602—TIS97

Miscellaneous
S15, S18—DPDT toggle switch
S16—SP3-position toggle switch
S17—SP10-position rotary switch
S19, S20—SPDT toggle switch
J6—3-terminal microphone connector
J7—BNC connector
M601—1-mA DC meter
OV601—LA10 overvoltage limiter (MCG Electronics)
F601—0.5-amp fuse

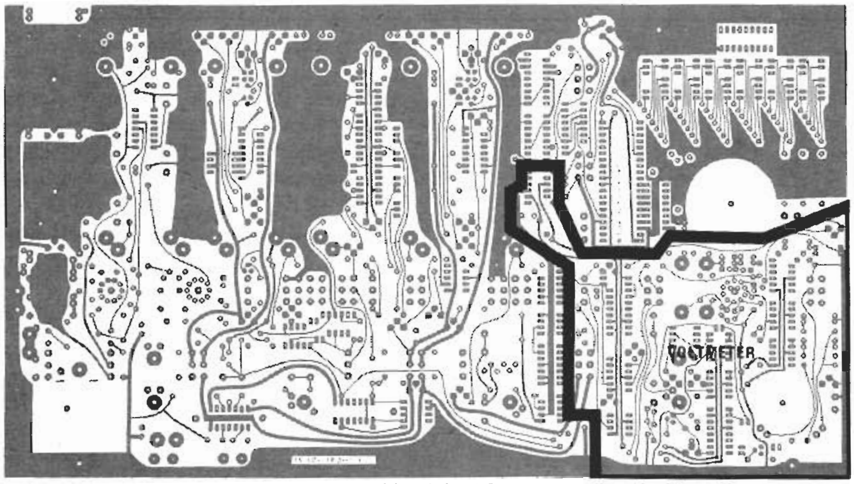


FIG. 11—FOIL PATTERN of the component side of the PC board overlayed with outlines of the area occupied by the voltmeter. The figure below shows the parts that are installed in this area.

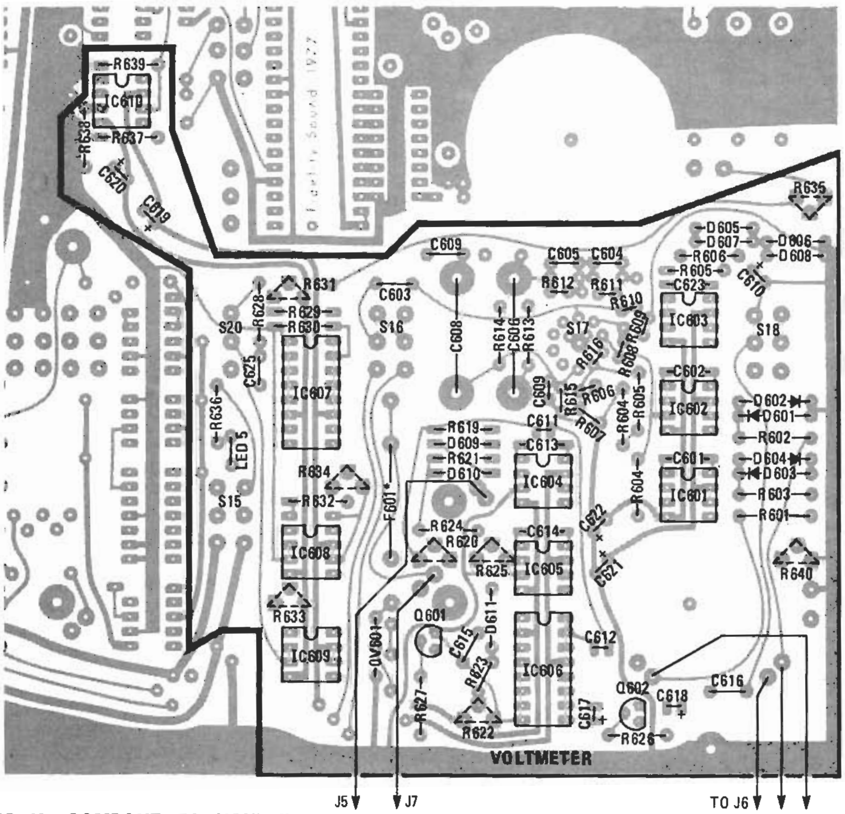


FIG. 12—COMPONENT LAYOUT diagram for the voltmeter. Parts used in this section are coded with numbers in the 600 series. See Fig. 10 for the schematic diagram.

The next task is to physically eliminate the lower portion of the function. To accomplish this, R631 applies a negative voltage to the inverting input of the output stage of IC607. This drives the log waveform up against a positive supply rail, which limits the transition or deflection. IC608 and its associated components provide proper polarity as well as gain and zero set.

Switch S20 selects the log signal, which was just described, or the linear signal, which is simply a bypass of the log converter. Resistor R635 converts the voltage signal to a current signal to drive the meter. A one-milliamp movement is quite satisfactory. Since audio measurements can sometimes be quite dynamic, it is preferable that the movement be critically damped.

IC609 is a very low-impedance output buffer. When connected in this manner, the impedance, looking back into the device, is that of the negative input, which is a virtual ground. IC609 is not really necessary but it does provide lower output impedance, isolation of the signal to the meter and better output protection than the circuit would have without it.

Overvoltage limiter OV601 limits the effect of the application of an external voltage; the same type optional unit is recommended here as was recommended for the other sections. Fuse F601 protects the overvoltage protector from excessive dissipation.

Assembly of this section is straightforward. Figure 11 shows the location of the voltmeter section with respect to the entire PC board and Fig. 12 shows the location of the components.

Counter section

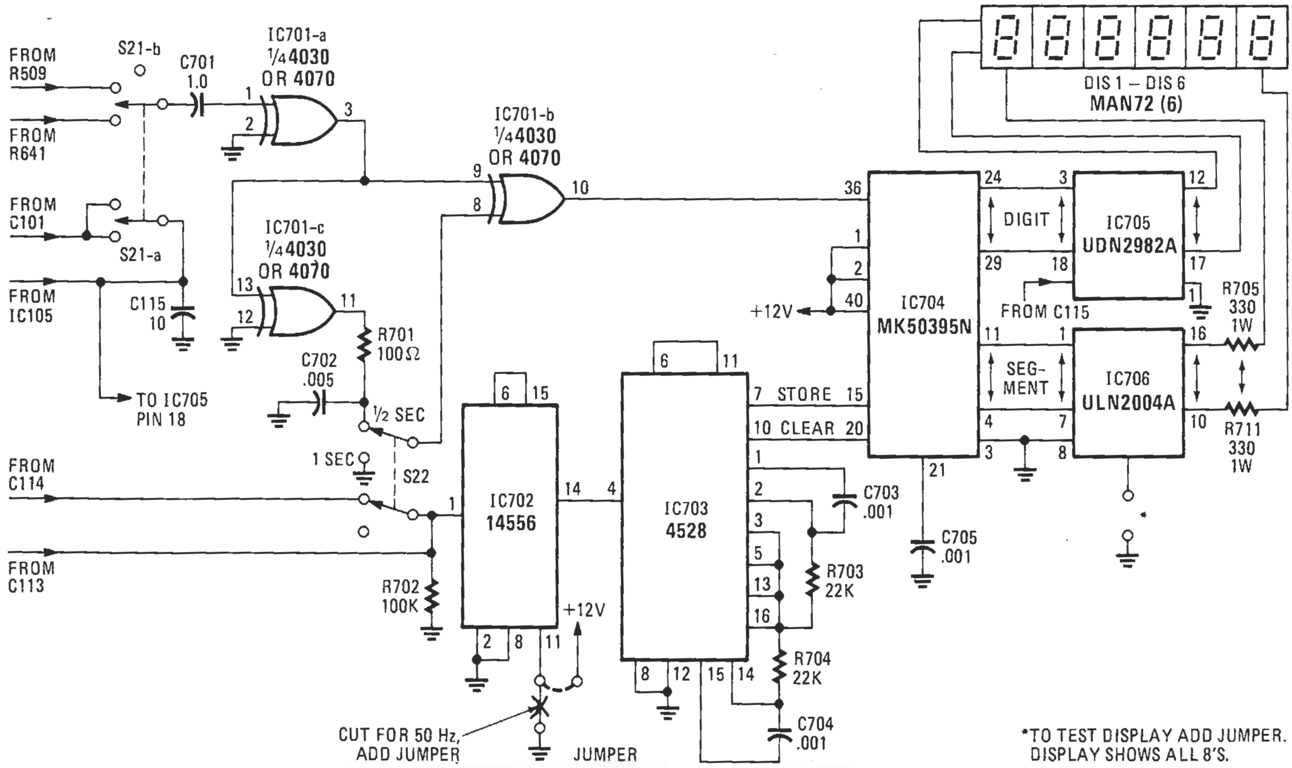
The schematic diagram of the frequency counter is shown in Fig. 13. Section b of S21 is the counter power switch; switch S21-a selects the signal to be counted. In one position the counter counts whatever signal is applied to the voltmeter. The voltmeter signal, however, can possibly be very low amplitude or altered in some way by the system under test so that it does not provide true and stable triggering of the counter. Therefore, S21-b can alternately select an internal signal from the audio sweep generator section. Stable counting of the sweep generator signal is thus always assured regardless of what happens to that signal as it is processed by the system under test.

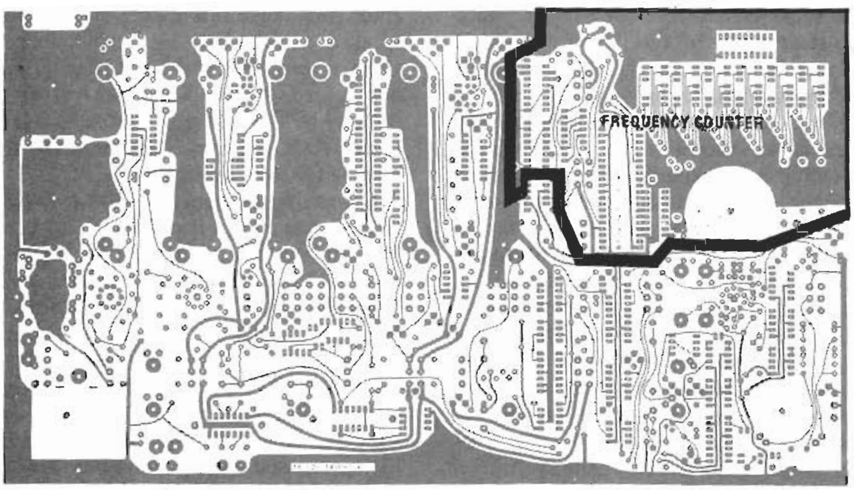
The counter timebase establishes the counting period or window. The counting circuit counts each cycle of the input waveform that occurs during the time the window is open. If the counting window is 1 second, the total number of cycles counted is the average frequency of the input signal during that second in Hz.

This system uses both 1-second and ½-second counting windows. The one-half second window is more useful for tracking a swept frequency.

The timebase, which establishes a counting window, is derived from the power line and selected by switch S22. Resistor R702 is a pull-down resistor for the input of counter IC702. With pin 11 connected to ground, IC702 divides by 60. With pin 11 connected to V+ (which for the counter section is 12V), it divides by 50. On the circuit board, pin 11 is connected to ground. To convert the unit to a 50-cycle power line, it is only necessary to cut a circuit on the back side of the board and install a short bus wire jumper between two pads, which are provided for this purpose. Cutting the circuit disconnects pin 11 from ground. Adding the jumper connects pin 11 to V+. The output of IC702 then is a pulse train whose frequency is either 1 or 2 times the power-line frequency.

IC703 is a dual one-shot. For each input pulse, IC703 provides a negative going store pulse. This store pulse is about 10-μs wide. This time is determined by C703 and R703. As the store pulse terminates, it triggers the second one-shot, which produces a positive going clear pulse. The width of the clear pulse is determined by C704 and R704 and is also about 10 μs. The counting window then is from the end of the clear pulse to the beginning of the next store pulse. These two pulses shorten the counting window by about 25 μs, inducing an error of 25 parts-per-million for a one-second update time or 50 parts-per-million for a half-second update. This error is insignificant in comparison to the normal power-line frequency variation. Line frequency er-





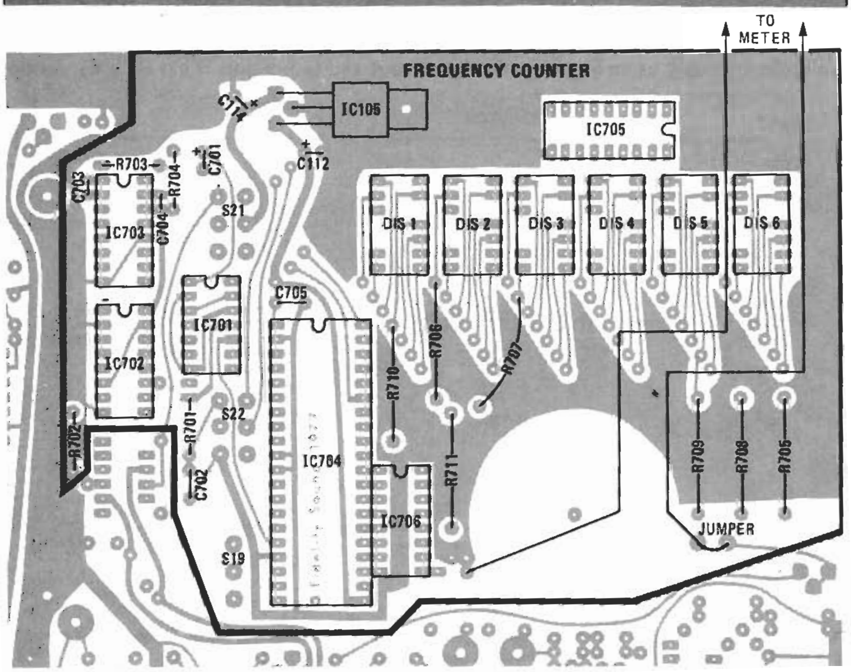


FIG. 13 (above)—THE SCHEMATIC of the digital frequency meter. FIG. 14 (left)—THE AREA OCCUPIED by the frequency meter. FIG. 15 (lower left)—LAYOUT OF PARTS in the frequency counter. IC105 at top center is a voltage regulator that is a part of the power supply.

COUNTER PARTS LIST

R701—100 ohms, ¼ watt, 5% or better
R702—100,000 ohms, ¼ watt, 5% or better
R703, R704—22,000 ohms, ¼ watt, 5% or better
R705-R711—330 ohms, 1 watt, 5% or better
C701—1 μF electrolytic, 16 volts or higher
C702—.005 μF disc ceramic
C703-C705—.001 μF disc ceramic
IC701—4030 or 4070
IC702—4566
IC703—4528
IC704—MK50395
IC705—UDN2982 (Sprague)
IC706—ULN2004 (Sprague)
DIS1-DIS6—MAN72 7-segment display
S21—DP3-position toggle switch
S22—DPDT toggle switch

The following are available from FSI, 1894 Commercenter W., No. 105, San Bernardino, CA 92408: Complete Kit, $495.00; cabinet and circuit board, $115.00. Set of semiconductors, $195.00; seven slide pots with knobs, $17.00, set of trimmers including four multiturn pots, $17.00.

California residents add state and local taxes as applicable.

rors are a fraction of a percent. The overall result is a timebase-caused error in the fourth or fifth significant figure of the displayed frequency.

IC701 is a quad exclusive OR gate connected as a frequency doubler. Gate

continued on page 105

AUDIO TEST STATION

continued from page 63

IC701-a is simply an isolator. With S22 in its lower position the timebase frequency is the same as the power-line frequency, which produces one window per second. At the same time pin 8 of IC701 is grounded, which yields an output pulse train of the same frequency as the input signal.

Gate IC701-c, R701 and C702 form the delay circuit. IC701-c provides both isolation and a small amount of delay and R701 and C702 provide additional delay. When S702 is in the one half-second update position, the input signal first arrives at pin 9 of gate A. Pin 8 is still

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low, the exclusive OR function is satisfied and pin 10 goes high. A short time later, the signal arrives at pin 8 and pin 10 goes low. Sometime later the input pulse ends. At this time, pin 9 goes low and pin 8 is still high. The exclusive OR is once again satisfied and pin 10 goes high again. A short time later the pulse at pin 8 also ends and pin 10 again goes low. The result is two narrow pulses for each single input pulse. The pulses occur at the beginning and end of the input pulse.

As the input frequency reaches about 300 kHz, the spacing betwen the two pulses reaches zero and the result is an output pulse the same frequency as the input. This is the only reason for having a one second update at all. That is, to extend the usefulness of the overall counter into this frequency range. If it was only to be used at lower frequencies, the one-second window need never be used.

IC704 is the actual counting element. When the clear input pin is low, IC704 counts the input at pin 36. When the store input pin is low, the number in the counter is transferred continuously to the display. When the clear input is high, the counter is reset. It starts over again at zero when the clear goes low. When the store input goes high, whatever number was last displayed remains displayed until the store goes low again. Since the store input only goes low for a brief period of time at the end of each counting window, the display shows the number the counter contained at the end of the previous window. Immediately after transferring the number present in the counter at the end of the period, the clear input goes high briefly, which initiates a new period.

The display is multiplexed. That is, only one digit is actually displayed at a time. IC705 provides power to the anode and IC706 provides ground for the cathode. Capacitor C705 establishes the scan frequency. IC704 through IC706 power each digit sequentially in seven segment format. If the LED's were continually powered, each number in the counter would flash sequentially on all six digits simultaneously. IC704 provides the digit scan through IC705. This simply powers each LED display sequentially. Therefore, a particular digit is only powered when the corresponding number from IC706 is to be displayed.

IC705 provides the voltage gain as well as a current gain. This allows IC704 to power the display from the unregulated supply. R705 and R701 provide current limiting for the displays.

Figure 14 shows the location of the Frequency Counter section and Fig. 15 shows the location of the components.

This wraps up the construction phase of the Audio Test Station. Next month we will go into calibration, tests and applications of this valuable addition to your bench. **R-E**