gain control IC
for audio signal processing

As a multi-purposed IC, the NE570 analog compandor fulfills many audio processing needs

Two ICs recently introduced by Signetics, the NE570 and NE571, permit the design of efficient and practical audio-signal control functions with a minimum overall parts count. These devices are primarily designed to act as compandors; the complementary processes of compression and expansion.[1,2] They are both dual-channel ICs and either portion can be used individually as a compandor. However, as will be seen in this article they are also well suited to a variety of other tasks useful to the amateur.

basic device operation

Each channel of the 570 and 571 consists of the functional components shown in **fig. 1A**. Packaged in a 16-pin DIP, the only items common to the two signal channels are the power supply, ground connections, and an internal 1.8 volt bias regulator.

The three principal components of each section are a ΔG cell, full-wave rectifier, and an output amplifier. The ΔG cell is used to control the gain over a range greater than 80 dB. The control voltage for this cell is generated by rectifying an input signal (RECT IN). The final output is then developed by the buffered output amplifier from the scaled signal current supplied by the ΔG cell. The 570 and 571 are identical electrically, but the 570 is selected for lower inherent distortion and a higher supply voltage range.

The ΔG cell, as shown in **fig. 1B**, consists of an op amp, A1, and transistor pairs Q1-Q2 and Q3-Q4. The input signal is first converted by R2 into a current that drives A1. The feedback for this op amp is via the transistor pair Q1-Q2. Therefore, the amount of current in this pair is the same as the current through R2. In addition to driving Q1-Q2, the op amp is also connected to Q3-Q4. Unlike Q1-Q2, this transistor pair does not have a constant-current source. By scaling the Q3-Q4 emitter current, their output is a linear product of the input signal from A1 and the scaled current. This circuit is a linearized transconductance multiplier[4-7] which cancels the inherent non-linearity and temperature sensitivity of the differential pairs, greatly enhancing the usefulness of this gain-control technique.

The rectifier portion consists of op amp, A2, class-B transistors Q5-Q6, a pnp current mirror Q7, and an npn current mirror Q9. When rectifying a signal at the RECT IN terminal, Q5 and Q6 produce pulses of current proportional to the positive and negative input signal swings. The output current of Q6 is used directly, while the Q5 current is mirrored by Q7. Thus, the drive to Q9 is a positive going, full-wave rectified pulsating dc. These pulses are filtered by an external smoothing capacitor attached to the C_{RECT} terminal.

The output stage is a simple inverting op amp similar in performance to a 741. Various options are possible by use of either R3, external input, or feedback resistors. The overall circuit gain is unity (ΔG IN to OUT), with R3 connected as a feedback resistor and 70 μA rectifier current into Q9.

In addition, the THD TRIM terminal allows a small offset to be introduced into the ΔG cell to null its distortion. The two input op amps (A1 and A2) are connected to the internal 1.8 volt regulator. Each op-amp input should be capacitively coupled while the input impedance is determined by R2 or R1, respectively. Circuit operation is very stable and immune to power-supply variations. A single supply voltage from +6 to +18 volts (571) or +6 to +24 volts (570) can be used, though the following applications will use a +15 volt supply.

basic compandor circuits

The 570 and 571 can be quite simply connected for their basic functions of expansion and compression, as illustrated in **fig. 2**. These circuits will not be dealt with in great detail because most amateurs will probably be more interested in some of the other uses. Also, compandor operation is covered in detail in other literature.[1-3]

The gain through the expandor shown in **fig. 2A** is $1.43\ V_{IN}$, where V_{IN} is the *average* input voltage.

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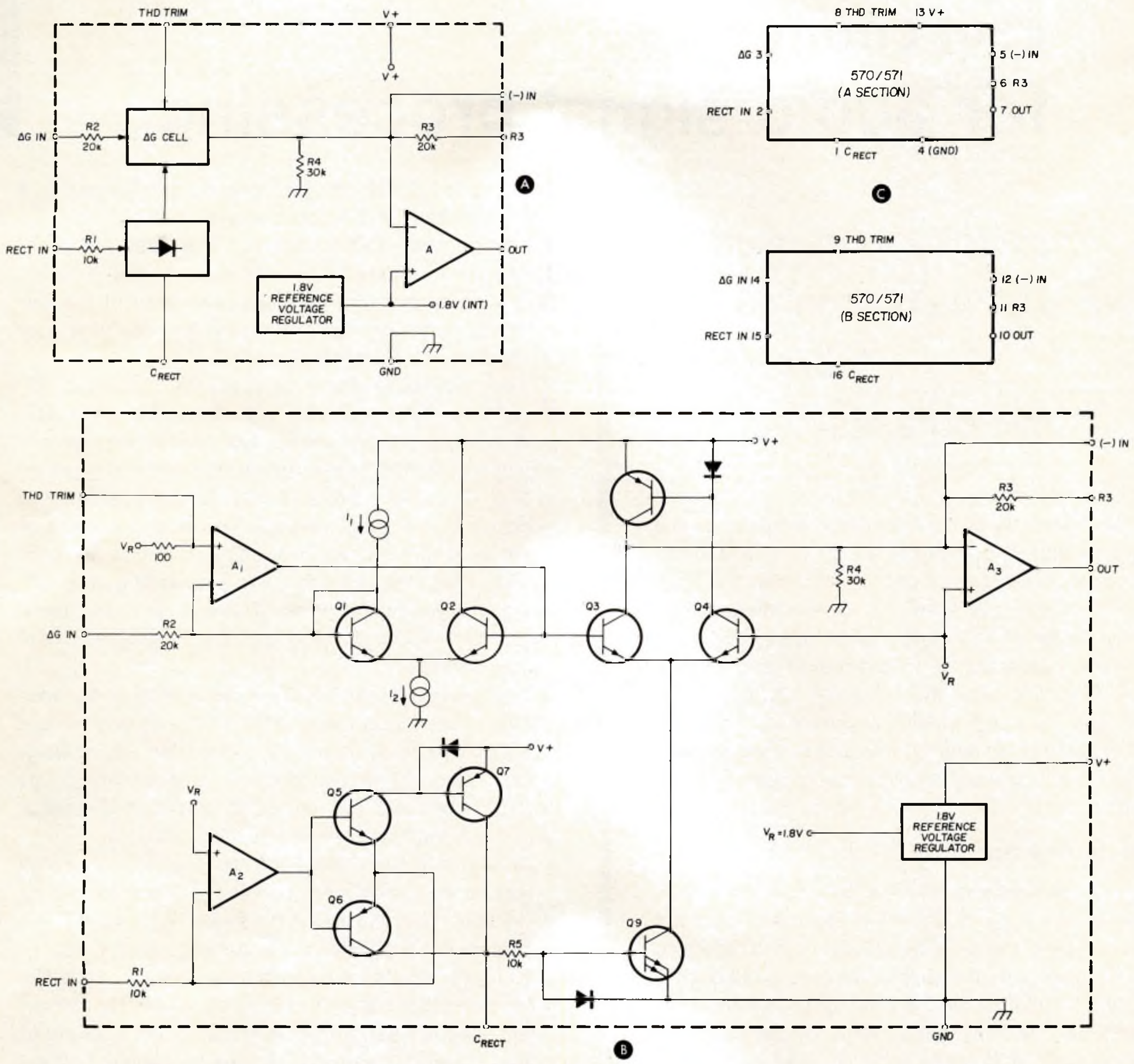


fig. 1. Functional diagram of the Signetics NE 570/571. A simplified schematic diagram of the device is shown in B. Constant current sources I_1 and I_2 feed transistor pair Q1-Q2.

The 570/571 circuit constants are set up such that unity gain occurs at an rms input level of 0.775 volts, or 0 dBm in 600-ohm systems. The C_{IN} and C_O are coupling capacitors, chosen for the desired low-frequency rolloff. C_{RECT} is selected for the desired time constant (10 ms) in conjunction with the internal 10-kilohm resistor (R5).

Resistors R_A, R_B, R_C and C_B are not essential to basic operation, but are desirable. R_B furnishes short-circuit protection for the output and capacitive load buffering, while R_A and R_C polarize C_{IN} and C_A. C_B is a power supply bypass, typically an aluminum electrolytic.

The compressor configuration in **fig. 2B** also has unity gain at 0.775 volt (rms) input, but, a complementary in/out characteristic. The main difference in this circuit is that the ΔG cell is connected as a feedback impedance via C_F, and the input is applied to R3 through C_{IN}. Bias for the output stage is set up by the RC-decoupling network, with the values shown appropriate for 15-volt power supply.

In general, the OUT terminal should be biased to one-half the supply voltage. Use of a 570 or 571 as a compandor is not limited to the gains shown, but may be extended to other ranges by use of additional components.

trimming techniques

Device performance can be enhanced by judicious trimming, as shown in **fig. 3**. Each technique is optional, and can be applied in any combination when the highest performance is desired. The most useful of the three methods is probably the THD trim, which minimizes the gain cell harmonic distortion. In this case, a small voltage (0 to 3 volts) is used to inject a current into the THD TRIM terminal through the 100k resistor. By biasing the rectifier terminal as shown, the inherent current flow in the rectifier is compensated for and permits better low-level signal tracking. Typically, the gain-control signal to A1 should not be reflected in the output. The control feedthrough trimmer will minimize that signal during periods of low input voltage.

applications

An interesting and versatile group of circuits, the gated or switched-mode amplifier, can be built from the 570/571. With the device controlled by external logic applied to the RECT IN input, the *on* gain is normally set to any value and the *off* attenuation can be in excess of 80 dB. Use of the 570 or 571 is advantageous in that all portions of the function can be performed entirely within the IC. Further, the on/off transition times can be set to a value determined by the time constant from C_{RECT}.

Fig. 4A is a logic controlled amplifier configured for a HIGH input to be on, and LOW off. When the control input is HIGH, CR1 is off and the current

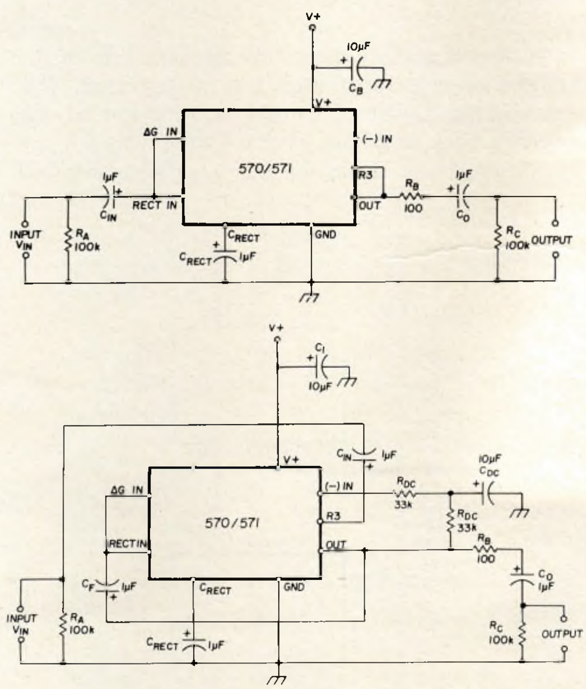


fig. 2. Schematic diagram of the devices connected as an expander, A, and a compressor, B. The voltage gain through the expander is $1.43 \cdot V_{IN}$, while for the compressor it is $\sqrt{0.7/V_{IN}}$. V_{IN} is the average input voltage.

developed by R_{GAIN} flows into the rectifier input, which turns on the ΔG cell allowing the signal to be amplified. R_{GAIN} can be selected for the desired *on* state gain, which is unity with a rectifier current of 70 μA. R1 and R3 also effect device gain, but R3 is selected basically for an optimum output bias of 7.5 Vdc. R1 can also be adjusted for gain, but as shown the value allows up to 3 volts rms input/output signal levels.

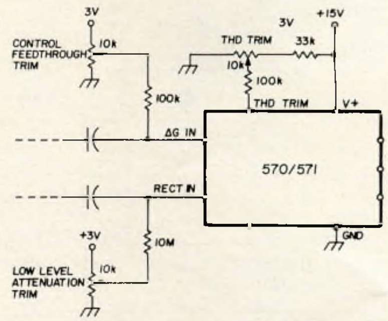


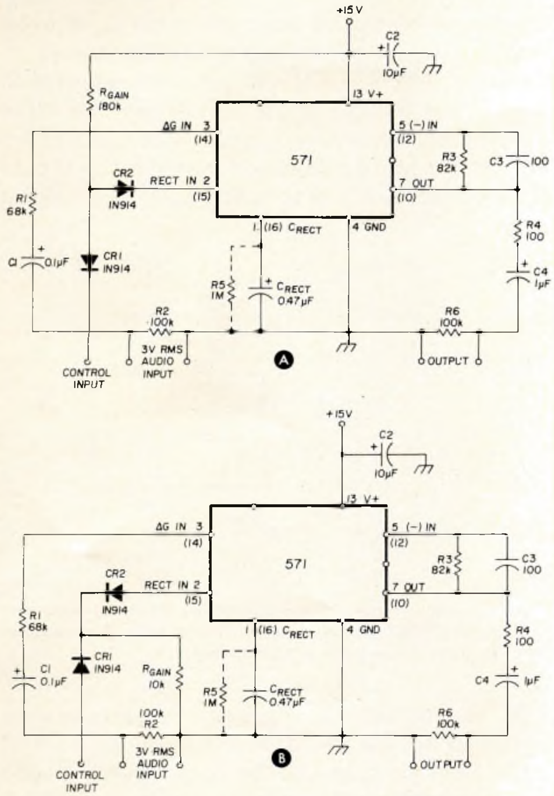
fig. 3. By applying the different trimming methods, distortion through the 570/571 can be reduced. Though each method is optional, they can be applied in any combination.

As can be seen in the control characteristics plotted in **fig. 4C**, the gain is unity (or its nominal value, if chosen otherwise) for control inputs greater than 3 volts. Switching is quite abrupt, with full attenuation being achieved at levels less than 1.5 volts. This narrow transition width and the nominal dc center of 1.8 volts allows direct control from CMOS, TTL, DTL, or other positive logic. The ultimate voltage of the HIGH state is non-critical, due to the 100-volt rating of CR1. Unfortunately, this circuit has one inherent weak point. Gain is sensitive to supply voltage due to the connection of R_{GAIN}. Thus, the supply voltage should be stable while choosing R_{GAIN} for 70 μA into the RECT IN terminal.

A companion circuit with complementary control characteristics is shown in **fig. 4B**. In this case, the gain is determined by the current developed through R_{GAIN} in conjunction with the internal voltage reference (1.8 V). With a low control input, the normal current will flow out through R_{GAIN}. When the control signal is high, CR1 is forward biased, interrupting the current flow. Therefore, the output will be attenuated since Q3-Q4 have been turned off.

Both circuits can be tailored for specific on-off transition times by selection of C_{RECT}. The time constant is simply $10k \cdot C_{RECT}$ (10 kilohms is the internal resistor). Thus, the audible switching effect can be smoothed, eliminating the transients produced by an asynchronous fast switch. The C_{RECT} value shown yields nominal times of 5 milliseconds.

Use of C_{RECT} in a switched amplifier of this type is



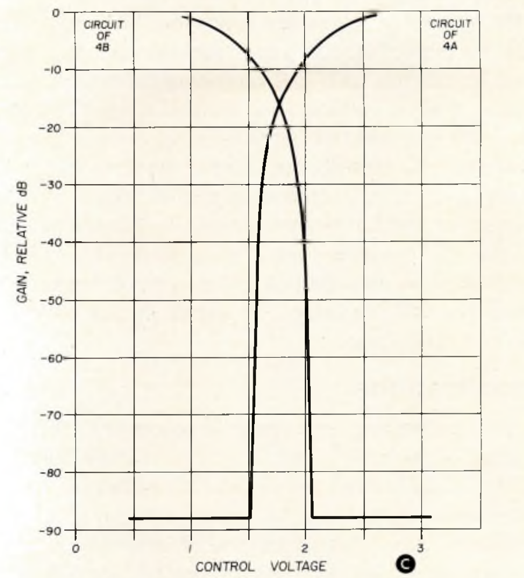


fig. 4. The logic controlled amplifiers can be configured to provide an output with either a HIGH or LOW input. This allows the amplifiers to be interfaced with many different logic types, TTL, CMOS, DTL, etc. The resistor R_{GAIN} should be selected to provide 70 μA at unity gain. Each amplifier has an on/off time determined by the time constant of C_{RECT} with internal 10k resistor.

optional and not absolutely necessary. However, to minimize noise pickup some capacitance will be found useful. Also, the ultimate *off* state attenuation will be limited to about 60 dB due to the internal bias current. This *feedthrough* error can be eliminated by connecting a 1 megohm resistor from C_{RECT} to ground to bleed away the error current. This allows attenuation of 80 dB or more.

Fig. 5 illustrates two sections of a 571 combined as a two-input multiplexer, for FSK or other uses. This circuit operation is similar to the others, but is biased and switched in a simpler manner. Gain of each *on* channel is unity, as determined by R_{GAIN}. The output of the B channel ΔG cell is summed with channel A by connecting the (−) IN terminals of the A and B sections. The respective channels are gated *off* by a low control logic input, which clamps the rectifier current, switching the ΔG cell off. For fsk or alternate channel use, the CONTROL A and CONTROL B signals should be complementary. Thus, the input is "instantaneously" switched between the **A** and **B** inputs.

Control signal suppression can be optimized with the CHOPPER NULL control, which trims the control signal component in the output. Suppression is better than 60 dB after trimming. Response time is quite fast, and is actually limited by the slew rate of the output op amp rather than the ΔG cell itself. This makes the switching interval a function of the signal's peak amplitude. For instance with a 4-volt peak amplitude signal the 0.5 V/μs slew rate will

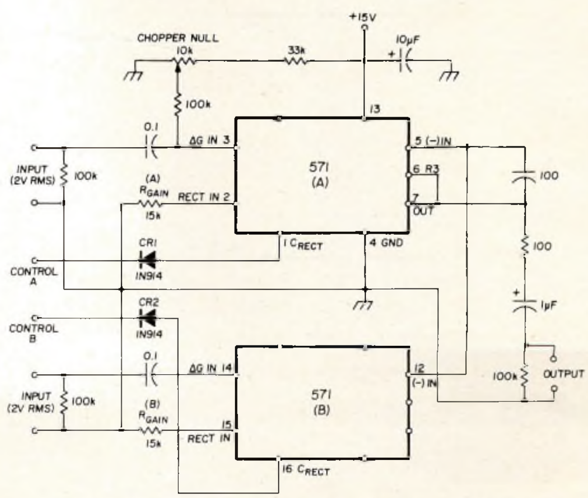


fig. 5. By providing complementary control signals, the FSK generator will switch between the two signal inputs. The outputs, when ON, are summed through the first operational amplifier.

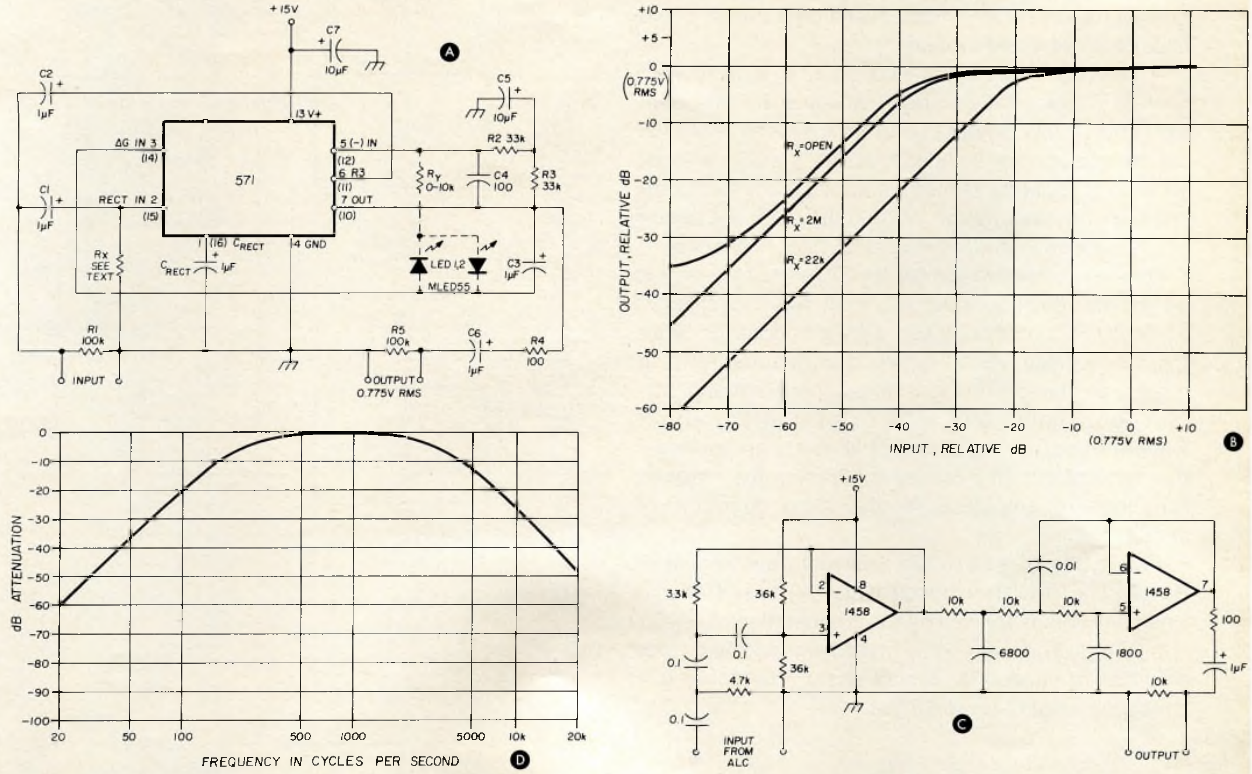


fig. 6. Schematic diagram of an NE570 connected as an automatic level control (A). The speech filter (C) is added to remove the undesired components created by the peak clipping. The attenuation curve (D), though not sharp, provides excellent rejection of the sharp pulse spikes generated by the clipping.

allow switching in 8 µs; lower amplitudes will be proportionally faster. Although the circuit is touted as a multiplexer, it can also be used as a summing switch, with both signals on at any given instant.

automatic level control

Automatic level control is a relatively common requirement in audio signal processing. A 571 automatic level control circuit can provide constant, high percentage modulation with varying input levels, yet without danger of overload if properly handled.[8] This circuit (**fig. 6**) is adapted from the 570/571 data sheet. There is one additional feature which may be useful, however: an optional resistor allows the threshold of level regulation to be varied. If R_X is left open, the circuit will have its widest range of gain control. As this resistor value is lowered, a larger input signal is required for full output. The general effect, for various R_X values, is shown in **fig. 6B**.

Since the 570/571 device operates on the principles of *average* level detection, it will easily saturate on large crest factor input levels such as speech, or large transient envelope changes. This condition is highly undesirable, but, fortunately its effect can be negated quite simply. Since the result of overshoots are peak-to-peak amplitudes in excess of the regulated output level, it follows that appropriate peak-level clipping can effectively control the overshoots. In this case, the rms output amplitude is 0.775 volt or 2.2 volts p-p. This particular level is conveniently clipped with a pair of reverse paralleled LEDs, which will limit to 3.2 volts p-p. The LEDs can be connected as shown in **fig. 6** with a series resistor R_Y, which is used to regulate the clipped amplitude.

This clipping technique, while not a requisite part of the automatic level control, greatly enhances its regulation with nonsinusoidal signals. Use of this circuit with speech inputs will necessitate the diodes, which will typically be clipping a good portion of the time. This, of course, adds audible distortion to the output. Therefore, a useful item with the ALC/clipper is a speech filter to remove the superfluous high and low frequencies. The filter will also greatly attenuate harmonic components generated by clipping.

The circuit in **fig. 6C** is a bandpass speech filter which uses a single IC. The circuit is simply a pair of cascaded Sallen and Key[10] highpass filters (3-pole Bessel type). The Bessel response is one of the poorest in terms of cutoff sharpness, but good from a pulse response standpoint. This feature is important to minimize amplitude overshoots which could occur with severely clipped inputs.

A common 1458 (dual 741) op amp is used with nearest 5 per cent component values for the filter elements. If low-power operation is desired, the 1458 can be replaced directly with a 358. If a 358 is used, 10 kilohm resistors should be added from each output terminal to common. With unity gain, the circuit can drive load impedances greater than 10 kilohms.

One very effective use for the 570 and 571 device is an amplitude-regulated RC sine wave oscillator. Typically, such circuits use a Wien bridge or other frequency-selective RC network, with some form of amplitude stabilization to maintain constant and correct loop gain, and also to guarantee output waveform purity. A 570 or 571 is nearly optimum for this type of circuit because it contains the required functions of amplifier, rectifier, and gain-control circuits.

Two types of sine wave oscillators are shown in **fig. 7**. The oscillator circuit (**fig. 7A**) based on the Wien network is formed by the combination of R1-C1 and R2-C2. This network is placed around the output amplifier of section A, which effectively makes it a bandpass amplifier resonant at

$$f = \frac{1}{2\pi RC}$$

With equal values of R and C, the input/output voltage ratio is 2 to 1.

To originate and sustain oscillations, the 571B section is used as an inverting amplifier with a nominal gain of 2. A slightly greater initial gain is established by the combination of R6, R7, and R8, which ensures startup. The B section ΔG cell is connected as a compressor, which regulates this stage's gain at the precise value required to maintain undistorted, stable amplitude oscillations.

There are two main steps taken to enhance flexibility of the circuit. A separate dc feedback path (R3, R4, C4) is used around the A stage, to remove value restrictions on R2 due to bias considerations. This allows R2 (R1) to range from 10k to 1 megohm without a major performance compromise. C1 and C2 have an even greater range, from 1 μF down to 100 pF. To minimize error due to strays, the lowest value should be used. With the values shown, the circuit is capable of reasonably low harmonic distortion. For example, 0.03 per cent distortion was measured at 1.6 kHz and THD (Total Harmonic Distortion) can generally be held below 0.1 per cent. This will vary according to the specific frequency, and the selected impedance of the Wien network. The low value of distortion is due to the light degree of ΔG cell regulation.

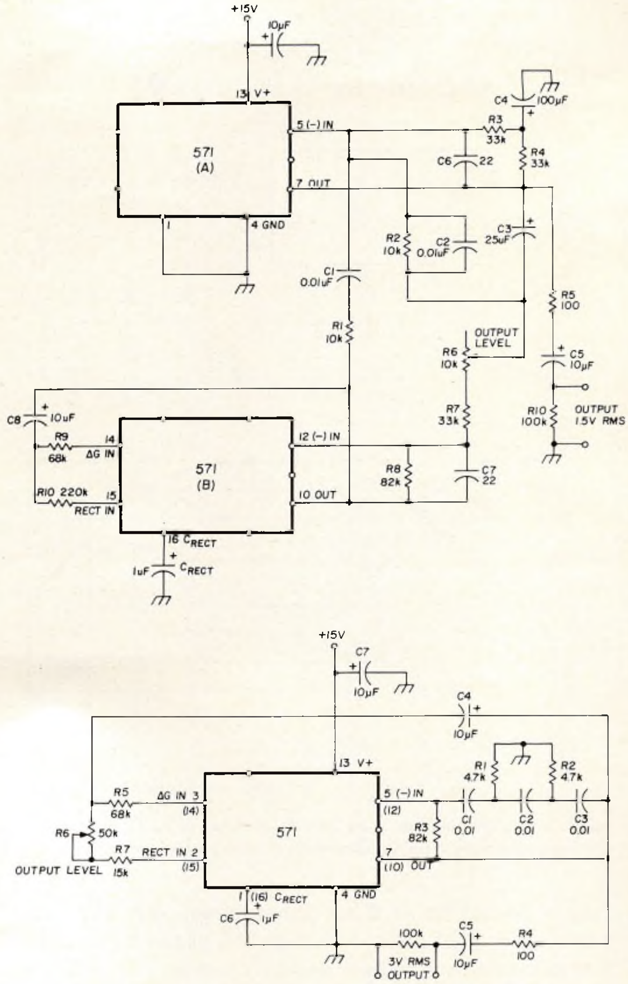


fig. 7. The NE570/571 can be connected as a sine-wave oscillator. The Wien bridge type oscillator is shown in A. For R = R1 = R2 and C = C1 = C2, the operating frequency is $1/2\pi RC$. Resistor R should be limited between 10k and 1 megohm with C between 1000 pF and 1μF. The normal frequency range can be varied from 10 Hz to 10 kHz. The phase-shift oscillator should be used to generate discrete frequencies only. Depending upon the selection of parts, the output frequency will be $1/2\pi RC\sqrt{3}$.

The circuit will operate as shown over the range from 10 Hz to 10 kHz. Below 10 Hz component size becomes impractical, and above 10 kHz slew limiting in the output amplifier causes distortion to rise. The circuit is useful as a fixed frequency oscillator, but can also be tuned if a matched dual pot is available for R1-R2. Output amplitude is set by R6, and is optimum at 1.5 volts rms output, from section A. If a higher output level is needed, section B output can also be used, at 3 volts.

The circuit of **fig. 7A** may be unduly complex for some uses, so an alternate and much simpler sinusoidal oscillator is shown in **fig. 7B**. This circuit is a form of phase-shift oscillator, similar to that described by Tobey, Graeme, and Huelsman.[9] A 571 is well suited for a phase-shift oscillator because it contains the necessary inverting amplifier to sustain oscilla tion. In the circuit shown, C1, C2, and C3 are the timing capacitors, while R1 and R2 are the resistors for the phase-shift network. R3 must be at least 12 times the R1-R2 value for adequate loop gain. AGC is provided by using the ΔG cell as a compressor.

This circuit is not suitable for tunable use. It should only be used as a spot frequency oscillator, by varying C1, C2 and C3. This is because R1 and R2 are related, by the design, to R3; in this specific case R3 cannot be variable because it is used to set the output dc bias point.

Although it uses a simple design, this circuit produces excellent results. At the frequency indicated, a laboratory test indicated a THD of 0.01 per cent at 3 volts output, which is remarkable in view of the circuit's simplicity. To take full advantage of this performance, an output buffer may be useful; for this you could simply use the remaining channel as a simple unity gain inverter.

conclusions

This discusssion has covered a few uses for a new and interesting chip. In the course of this article's preparation several other potential uses suggested themselves, such as phase comparators, phase-locked loops, voltage-tuned oscillators, and others. Unfortunately, space and time restrictions did not permit their complete examination.

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