INSTRUMENTATION & TECHNIQUES

A programmable rippled-noise generator for auditory research

LAWRENCE L. FETH and MICHAEL J. DONOHUE
Department of Audiology and Speech Sciences, Purdue University, Lafayette, Indiana 47907

This paper describes the use of two pseudorandom noise generators designed by Wolf and Bilger (1977) to generate repeatable samples of noise with sinusoidally rippled amplitude spectra. A description is given of the interface necessary to program the generator using a laboratory computer. Small modifications to the original generators are required. Sample output spectra are also shown.

Broadband noise signals with sinusoidally shaped amplitude spectra are of use in the study of auditory frequency selectivity (e.g., Houtgast, 1972; Wilson & Evans, 1971) and pitch perception (e.g., Atal, Schroeder, & Kuttruff, 1962; Bilsen, 1977). Such signals are also used in neurophysiological studies of the auditory system (Narins, Evans, Pick, & Wilson, 1979).[1]

The conventional way to generate a rippled-spectrum noise requires that a broadband white noise be added to a time-delayed version of itself to create an electrical interference pattern, which produces the spectral ripples. This result can be visualized by first considering only one sinusoidal component of the broadband noise. If the time-delay T is equal to the period of that component, then the addition of delayed and undelayed waveforms should produce an output with an amplitude twice as large as the original. The same should hold true whenever the delay is equal to an integral number of periods of the component. For very long delays the summed waveform may not be exactly twice the original because of the fluctuations in the noise.

To continue the visualization, consider the resulting waveform when the delay is equal to one-half the period of the component. The original is added to a delay waveform that, in effect, has been inverted. The result, in an ideal system, is perfect cancellation. Also, for any delay that is equal to an integral multiple of half-periods, there will be such cancellations. If the two branches of the delay-and-summing circuit have slightly unequal gains, the cancellation will be less than perfect. Also, noise fluctuations may degrade cancellations for very long delays.

For a given time delay, each of the components of a broadband noise encounters the delay-and-summing circuit. Because the circuit is linear for all practical

purposes, the output signal is the superposition of the output due to each individual component. The components that encounter a delay equal to an integral multiple of their period will be "reinforced" in the output signal. That is, peaks in the output spectrum will occur at frequencies equal to n/T, with $n = 0, 1, 2 \ldots$ ($n = 0$ corresponds to the peak at dc). Components for which the delay represents an integral multiple of half-periods fall into valleys in the output spectrum. It should be obvious that spectral peaks and valleys are periodic in frequency and that the valleys fall halfway between the peaks.

It is not as simple to visualize the fate of components for which the delay equals some fraction of a period other than one-half. Analysis of the delay-and-summing circuit indicates, however, that the spectral transitions from peak to valley and back to peak is sinusoid. A plot of the circuit's transfer function on linear coordinates would reveal the sinusoidal shape.

If one branch of the delay-and-summing circuit contains an inverter, spectral peaks and valleys can be interchanged. Figure 1 shows the results of such an inversion. The spectra in Figure 1 do not appear sinusoidal because the ordinate is a logarithmic (decibel) scale. A broadband phase shifter in place of the inverter allows movement of the spectral peaks and valleys (actually the whole spectrum) along the frequency axis without changing the spacing. An attenuator can be introduced to reduce the peak-to-valley difference should that option be necessary for the experimental procedure.

Either analog or digital delay lines may be used, but there are drawbacks. Analog delay lines are not readily available for long delays (tens of milliseconds). Both delay lines may have poor (i.e., irregular) response characteristics over the frequencies of interest.

DESIGN APPROACH

Our approach to this problem, like that of Narins et al. (1979), was to use two identical pseudorandom
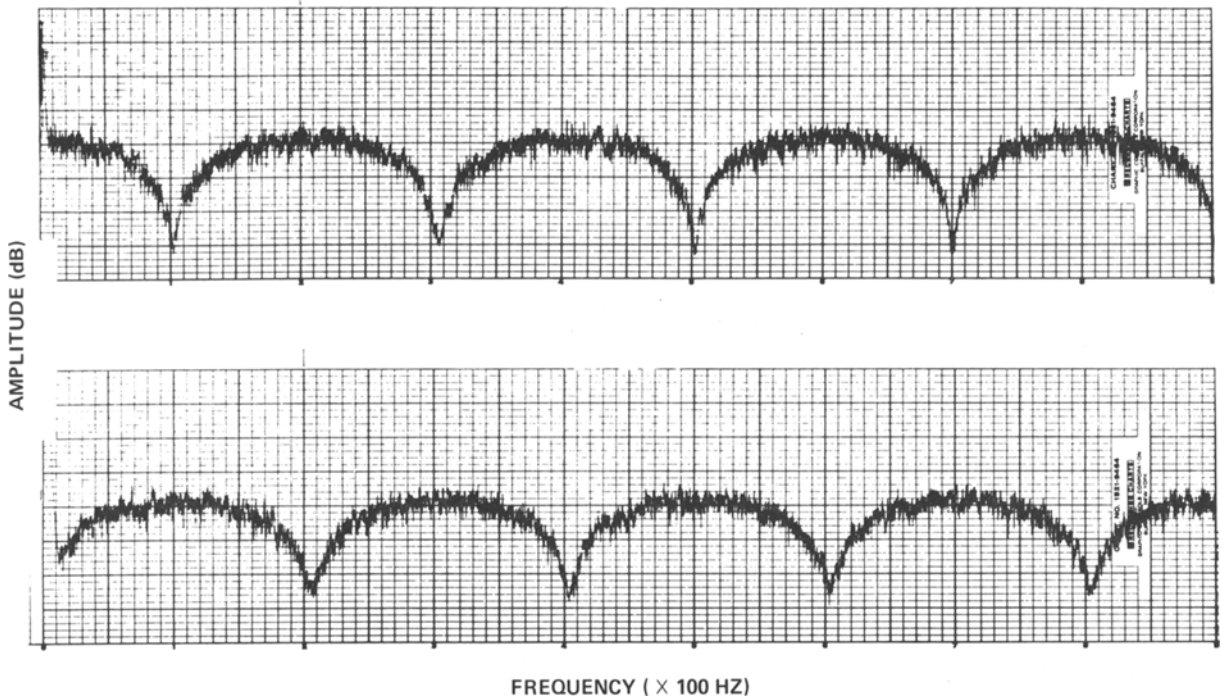
Figure 1. Amplitude spectra for the output of our rippled-noise generator. The ordinate is logarithmic (i.e., amplitude in decibels). The abscissa is linear in frequency. The time delay used to produce these spectra was 5 msec. The lower tracing shows the result of inverting one signal before combining the two generator outputs (see text for details).

shift-register noise generators with an offset in time. That is, rather than delay one branch of the circuit, we advanced one generator. Because they are identical, the two pseudorandom generators produce the same waveform if started from the same initial state and driven from a common clock. Our pseudorandom rippled-noise generator consisted of two of the generators described by Wolf and Bilger (1977).

The advance was accomplished by building a circuit which allows one generator to be driven by N clock pulses before the second generator receives any clock pulses. After the advance, both generators are driven from the single clock. The advance count N is delivered to the counter through a 16-bit latch circuit that is easily interfaced to either an 8-bit microcomputer or a 16-bit minicomputer. Figure 2 shows the interface, latch, and down-counter circuit.

For operation with an 8-bit microcomputer, the data lines are connected in parallel to the two data latches (U1-U2) and the load lines are connected to separate port enable lines. For a 16-bit machine, the data lines are run separately to the latches; the load lines can be tied together and given a common device code.

CIRCUIT DESCRIPTION

The circuit consists of an input latch (U1-U2), a down-counter (U3-U6) that counts down from N to zero, a zero-detection circuit (U7-U8), and an enable circuit for Clock 1 and for Clock 2. A reset circuit initializes both noise generators and the enable circuits for Clock 1 and Clock 2 and a test point that can be connected to a digital counter for direct measurement of the advance count.

The input latches (U1-U2) are enabled by a high level of their respective load lines. The least significant bits are loaded into U1 and the most significant bits are loaded into U2. The 16 data bits are then transferred to the down counters by a low level on the counter load lines. Part of inverter U12 converts a high level from a port enable line to the low level required by the counters.

A high level at the start input is converted to a low level by part of U12 and enables Clock 1 and the down counter. The clock pulses from the crystal-controlled clock advance Noise Generator 1 and the down counter simultaneously.

The down counter is triggered on the rising edge of the clock pulse. When all four stages of the counter have reached zero, the output of U7 goes low and enables the J-K flip-flop (U8). The flip-flop is a negative-edge triggered device; that is, it is triggered by the falling edge of the crystal clock. In this way the down counters reach zero on the rising edge of the Nth clock cycle and the flip-flop changes state on the falling edge of the Nth clock cycle. When U8 changes states it enables Clock 2. Both noise generators are then clocked simultaneously.

The down counters and the test point are enabled only when Clock 1 is being advanced and Clock 2 is disabled. A high level on the RESET line disables both Clock 1 and Clock 2, as well as initializing both noise
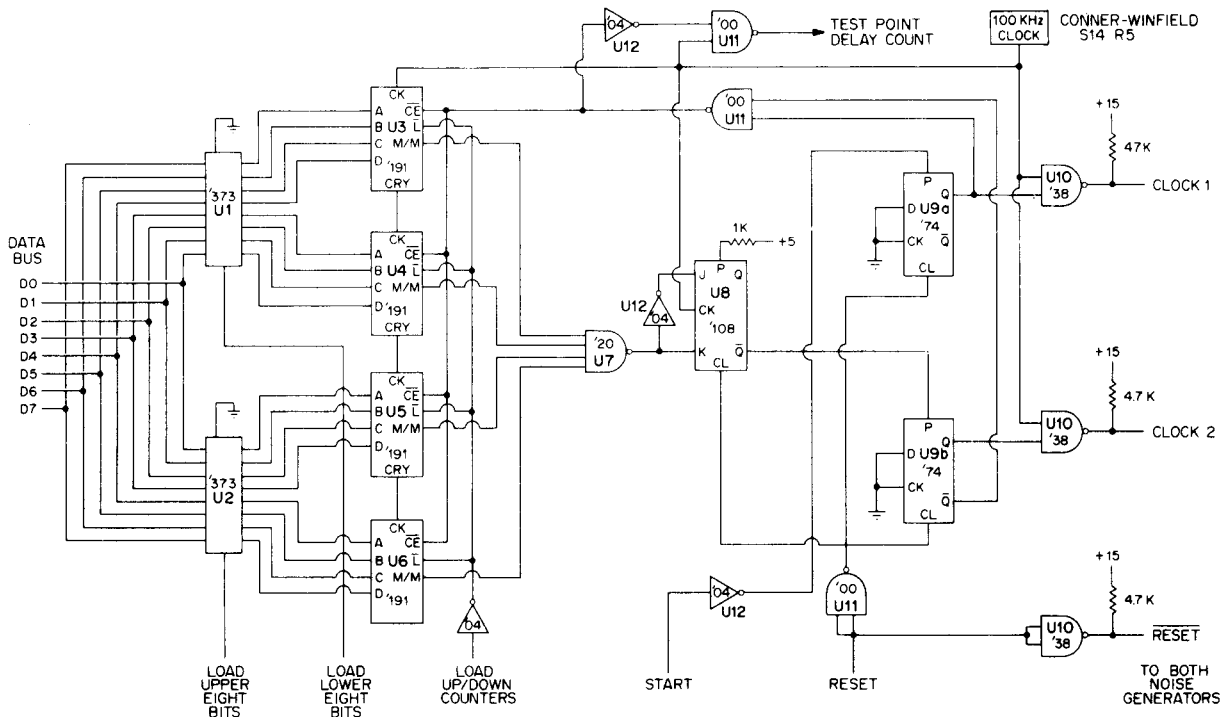
Figure 2. Interface for programming the advance of one pseudorandom noise generator over the other.
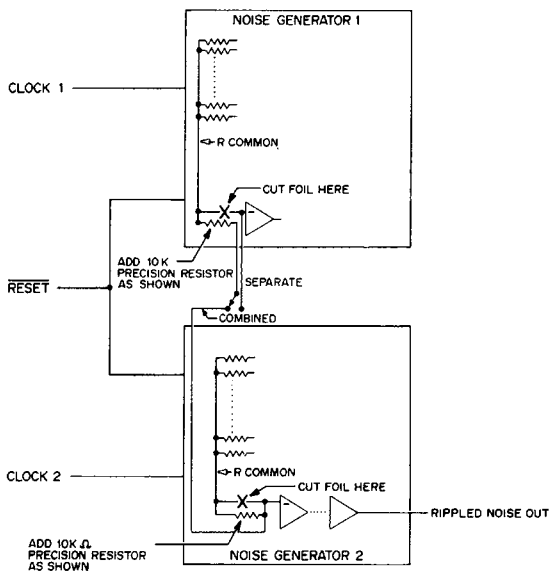

Figure 3. Modifications required in the original pseudorandom noise generators to allow for either a common or a separate output signal.

generators. This must be done each time the rippled-noise count is changed.

MODIFICATIONS TO NOISE GENERATORS

Operation of the rippled-noise generator requires a small modification of the noise generators (Figure 3).

The line from the resistor common must be cut where it goes into the inverting input of the low-pass filter (U6-Pin 5) on each noise generator board. The switch (SI) is added to provide either separate or combined outputs from the two noise generators. In the position shown in Figure 3, the pseudorandom square waveforms are added together through a resistive mixing circuit. The combined waveform is fed to the low-pass filter on Generator 2 (Pin 5 of V 6). A rippled noise then appears at the output of Generator 2.

In the other switch position, the two waveforms are available at separate output connectors, so that an inverter, phase-shifter, or attenuator may be introduced before the two waveforms are combined.

The internal clock on the sine-noise generators must be disabled by removing the jumper connecting the onboard clock (U7-Pin 10) to the clock inputs of the shift registers (U1-U4). The external clock from the ripple-noise interface (Clock 1 and Clock 2) is then connected to the clock inputs of the shift resistor of Noise Generator 1 and Noise Generator 2, respectively.

DISCUSSION

The rippled spectra shown in Figure 1 were produced by our programmable pseudorandom generator, not by a delay-and-add circuit. The peak to valley ratio is at least 30 dB and can be improved slightly with some fine tuning of the noise generators and our mixer circuit. Since the time required to move from one advance (delay) time to another is approximately equal to the new time interval, this generator can be used in studies that

require rapid changes in the ripple spacing. For example, we have programmed simple melodies with the generator for pitch demonstrations.

REFERENCES

ATAL, B. S., SCHROEDER, M. R., & KUTTRUFF, K. H. Perception of coloration in filtered gaussian noise: Short-term spectral analysis by the ear. *Proceedings of the Fourth International Congress on Acoustics (Copenhagen)*, 1962, **H31**, 1-4.

BILSEN, F. A. Pitch of noise signals: Evidence for a "central spectrum." *Journal of the Acoustical Society of America*, 1977, **61**, 150-159.

HOUTGAST, T. Psychophysical experiments on grating acuity. In *Symposium on hearing theory*. Eindhoven, The Netherlands: IPO, 1972.

NARINS, P. M., EVANS, E. F., PICK, G. F., & WILSON, J. P. A comb-filtered noise generator for use in auditory neurophysiological and psychophysical experiments. *IEEE Transactions on Biomedical Engineering*, 1979, **BME-26**, 43-47.

WILSON, J. P., & EVANS, E. F. Grating acuity of the ear: Psychophysical and neurophysiological measures of frequency resolving power. *Proceedings of the Seventh International Congress on Acoustics (Budapest)*, 1971, **7**, 397-400.

WOLF, R. V., & BILGER, R. C. A digital noise and sine wave generator. *Behavior Research Methods & Instrumentation*, 1977, **9**, 345-348.

NOTE

1. The original version of the generator was designed and built at about the time that the circuit described by Narins et al. (1979) was published. We were unaware of Narins' work, however, until the spring of 1980. Both generators operate on the same principle. Narins chose a higher clock frequency, which is apparently necessary for physiological frequency selectivity studies. The generator described in this paper is programmable by a laboratory computer. Features of both generators can be combined easily in any generator the reader may wish to build.