a test setup. Ground currents can slow the settling time of voltage steps applied to a test circuit or the output voltage to be measured.

Acknowledgment

The author wishes to thank B. A. Bell and T. M. Souders of the Electrosystems Division, NBS, for their very helpful suggestions in preparing this manuscript.

References

[1] A. G. Perrey and H. K. Schoenwetter, "A Schottky diode bridge sampling gate," Nat. Bur. Stand (U.S.) Tech. Note 1121, May 1980.

[2] J. A. Whealler, "A new technique for testing settling time in a production environment," in *Dig. 1981 Int. Test Conf.* (Philadelphia, PA, Oct. 1981).

[3] Tektronix, Inc., Instruction manual for "7S14 Dual Trace Delayed Sweep Sampler" (Beaverton, OR, 1973).

[4] Tektronix, Inc., "DAC measurements: The sampling oscilloscope approach," Sampling Appl. Note 42K1.0 (Beaverton, OR, 1977).

[5] T. M. Souders, D. R. Flach, and B. A. Bell, "A calibration service for analog-to-digital and digial-to-analog converters," Nat. Bur. Stand. (U.S.) Tech. Note 1145, July 1981.

[6] Teledyne Philbrick, "Specifying and testing digital-to-analog converters," Bull. AN-25 (Dedham, MA 1976).

[7] D. H. Sheingold, Ed., *Analog—Digital Conversion Notes.* Norwood, MA: Analog Devices, Inc., 1980, pp. 207–211.

[8] T. M. Souders, D. R. Flach, and T. C. Wong, "An automatic test set for the dynamic characterization of A/D converters," this issue, pp. 180–186.

[9] Burr-Brown, "DAC 63, ultra-high speed digital-to-analog converter," specifications and applications information (Tucson, AZ., 1981).

[10] D. B. Bruck, *Data Conversion Handbook.* Burlington, MA: 1974, pp. 2-51, 2-52.

[11] Precision Monolithics, Inc., "DAC-10, 10-bit high speed multiplying D/A converter," specifications and applications information (Santa Clara, CA, 1980).

[12] J. R. Andrews *et al.* "Survey of present waveform sampling system limitations," NBS Rep. 10731, Feb. 1972.

[13] J. R. Andrews, B. A. Bell, N. S. Nahman, and E. E. Baldwin, "Reference waveform flat pulse generator," this issue, pp. 27–32.

Reference Waveform Flat Pulse Generator

JAMES R. ANDREWS, SENIOR MEMBER, IEEE, BARRY A. BELL, MEMBER, IEEE, NORRIS S. NAHMAN, FELLOW, IEEE, AND EUGENE E. BALDWIN, MEMBER, IEEE

Abstract—The NBS Reference Flat Pulse Generator (RFPG) is used to transfer dc voltage and resistance standards to the nanosecond domain. It provides a step amplitude of 1.000 V (open circuit) from a source impedance of 50.0 Ω. The transition duration is 600 ps, and all perturbations are damped out to less than ±10 mV within 5 ns. It can also be used as a time interval transfer standard.

Introduction

THE NEED FOR a reference step-like generator has become increasingly important for accurately characterizing the step response of oscilloscopes, transient recorders, and fast sampling channels of digital measurement instruments and automatic test equipment (ATE). By comparing the measurement waveforms and/or data with those of a reference waveform, the fidelity with which the measurement system can acquire the waveform can be determined. The dynamic performance of a measurement system is often desired, together with its ability to measure dc or steady-state quantities.

From a practical point of view, it is often desirable to be able to calibrate the dc voltage levels of the measurement system, as well as any time-dependent parameters. For this reason a voltage step function, whose beginning and ending dc levels are calibratable, has considerable merit as a reference standard, particularly if the transition between these two levels is also well behaved and predictable. Such was the purpose for developing the Reference Flat Pulse Generator (RFPG) at NBS.

Available Waveform

In theoretical time-domain studies the ideal unit step generator is commonly used as a driving source. However, in experimental work the ideal generator must be replaced by a physically realizable one, i.e., a step-like transition generator.

The voltage step-like transition generator produces a noninstantaneous voltage transition from an initial voltage level to a final one. An equivalent circuit for the voltage transition generator may be characterized in terms of the step response $E_g(s)$ of a four-terminal network having a voltage transfer
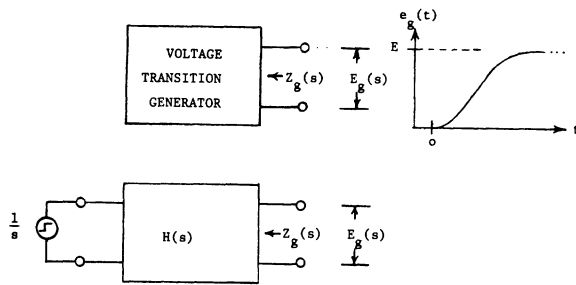
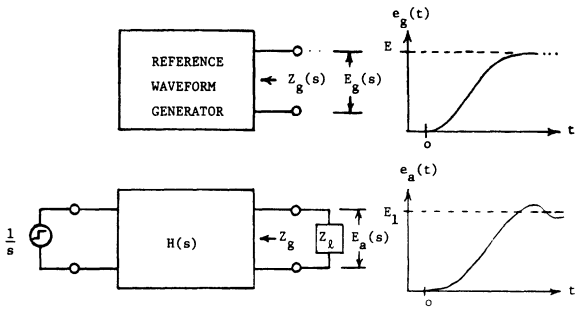Fig. 1.   An equivalent circuit for a voltage transition generator.

Fig. 2.   The available waveform, $e_a(t)$, across a load $Z_l(s)$ by the reference waveform generator.

function $H(s)$ and an output impedance $Z_g(s)$ (Fig. 1 [1]). When a transition generator is connected to a load $Z_l(s)$, the "available" transition waveform across the load is given by

$$E_a(s) = \frac{H(s)Z_l(s)}{s[Z_g(s) + Z_l(s)]} \qquad (1)$$

$$= \frac{E_g(s)Z_l(s)}{Z_g(s) + Z_l(s)}. \qquad (2)$$

Consequently, the "available" waveform at the load is dependent upon the open-circuit generator waveform $E_g(s)$, the generator output impedance $Z_g(s)$, and the load impedance $Z_l(s)$ (Fig. 2). If the load impedance $Z_l(s)$ is not equal to the generator impedance $Z_g(s)$, $e_a(t)$ will not have the same waveshape as $e_g(t)$. When $Z_l(s) = Z_g(s)$, then $e_a(t) = 0.5$ $e_g(t)$.

A reference waveform generator is one that produces a known available waveform across a specified load impedance. By "known" it is meant that a physically based model itself will always be an estimate due to the limitations of theory and measurements; consequently, the known available waveform will also be an estimate based upon the best knowledge available at the the time [1].

OTHER REFERENCE STEP GENERATORS

The fastest step generators to date have used either mercury-wetted mechanical switches or tunnel diodes. High voltage pulses (tens to hundreds of volts) can be generated using mercury switches. In addition, transition times of the order of 40 ps can be obtained using a miniature mercury-wetted switch [2]. Disadvantages of mercury switches are: low repetition rates (60 Hz), limited life times, and pulse amplitude instabilities of the order of a percent. Another more subtle

problem stems from the use of transmission lines in the typical mercury switch pulse generator design. The skin effect losses in the coaxial cable charge line cause a definite distortion of the pulse topline.

Commercial tunnel diode pulse generators exist with transition durations of the order of 20 ps. These generators provide extremely fast transitions but exhibit some waveform distortions due to trigger ramp, etc. The cleanest commercial TD step generator was described in detail by George Frye at the 1968 CPEM Conference [3]. The major limitation in using tunnel diodes is the long-term sag in their output voltage due to internal heating of the junction [3]. Thus they cannot be used for accurate calibration transfer from dc.

Other examples of reference waveform generators are the NBS reference waveform generators based upon transmission line dielectric loss produced by liquid dielectrics exhibiting a Debye relaxation process [1], [4], [5]. These generators employ dielectric loss to realize uniform transmission line filters. The filters provide a reflection-free low-pass bandwidth limiting process and, in turn, a Gaussian-like transition waveform upon being driven by a tunnel diode transition generator. Using a tunnel diode transition generator having about a 20-ps transition duration, three filter assemblies provide available waveforms having 50-, 100-, and 200-ps transition durations across a 50-Ω load. These generators can be used to evaluate the high-speed, leading-edge transient performance of oscilloscopes and other waveform recording devices.

FLAT PULSE GENERATOR CONCEPT

A flat pulse generator is needed to bridge the gap between dc voltage standards and the extremely fast transient generators such as tunnel diodes. In 1971, Andrews proposed and built a simple circuit which showed that it would be possible to build a generator to transfer dc voltage to the nanosecond region. The simple block diagram shown in Fig. 3 illustrates the principle.

Assume the load R_l is disconnected. Initially, the switch $S1$ is open. Current I_{dc} from the current source flows through the ideal diode $CR1$ and the resistor R_g. Thus the open circuit voltage developed across the output terminals is simply $V_{oc} = I_{dc} \times R_g$. The output impedance Z_g, seen looking back into the output terminals, is simply R_g since a current source, by definition, is an infinite impedance. Now at $t = t_0$, the switch closes. The anode of diode $CR1$ is pulled down to $-V$. The diode is now reverse biased and ceases to conduct, thus disconnecting the output from the current source and the switch. All of the current I_{dc} now passes through the switch to $-V$. The open circuit output voltage is now precisely 0 V. The output impedance is still R_g. Thus we have a generator which switches between two known output voltages ($I_{dc} \times R_g$ and 0 V) with a constant output impedance R_g.

In 1981, NBS built a prototype RFPG implementing the above concept. Fig. 4 shows the basic schematic diagram for the output stage of the RFPG. $Q4$ functions as the dc current source I_{dc}. $Q2$ is the switch $S1$. $Q1$ and $Q2$ form an emitter-coupled, differential switch pair. They are driven in push–pull by the complementary outputs of an ECL gate. $Q3$ is the
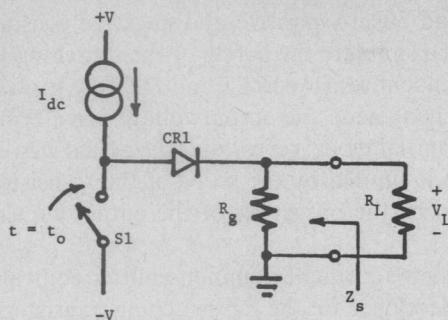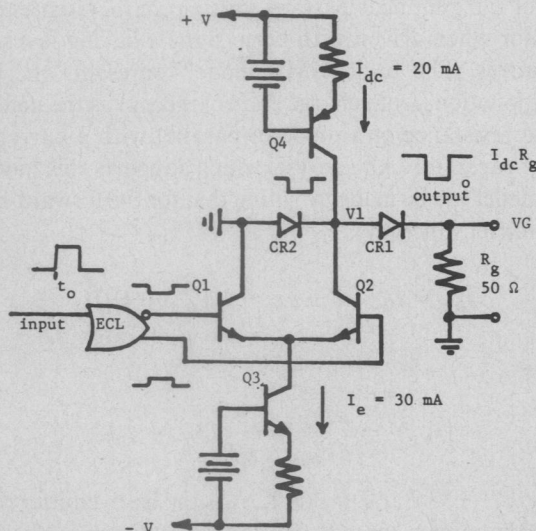
Fig. 3. Basic schematic diagram of flat pulse generator.

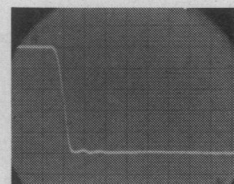Fig. 4. Output stage of the RFPG. *VG* is the open circuit generator voltage.

Fig. 5. NBS RFPG.

Fig. 6. NBS RFPG transition waveform. Vertical scale is 100 mV/div. Horizontal scale is 1 ns/div. This is the observed waveform across an external 50-Ω load.

current source for the emitters of $Q1$ and $Q2$. I_e must be greater than I_{dc}. Initially the base of $Q1$ is an ECL "1" (-0.9 V) and the base of $Q2$ is an ECL "0" (-1.7 V). Thus $Q2$ is off and $Q1$ is on, conducting all the emitter current I_e to ground. The current I_{dc} from $Q4$ all passes through $CR1$ to R_g creating an open circuit output of 1.00 V (20.0 mA $\times$ 50.0 Ω). $CR1$ is conducting while $CR2$ is reverse biased and nonconducting. At $t = t_0$ the input to the ECL gate is switched. The ECL complementary outputs rapidly switch $Q1$ off and $Q2$ on. $Q2$ now conducts all of the 30-mA emitter current I_e. Thus the collector of $Q2$ tries to pull 30 mA from the node connecting $CR1$, $CR2$, $Q2$, and $Q4$. The only way for the nodal currents to balance is for $Q4$ to supply 20 mA, $CR1$ to become nonconducting, and $CR2$ to turn on and supply 10 mA. Hence, we have accomplished the desired function of turning off $CR1$ and disconnecting I_{dc} from the output. In actual practice, only the negative going transition from 1 to 0 V is considered as the reference step transition. It is far easier to control the transient behavior by turning off $CR1$ than when turning it back on.

NBS RFPG

The NBS RFPG is shown in Fig. 5. Fig. 6 shows the reference transition waveform, i.e., the observed waveform across a 50-Ω load, as observed on a sampling oscilloscope. The output

transition waveform into 50 Ω is a negative going step from 500 mV (1.000-V open circuit) to 0.000 V with a 90- to 10-percent transition duration of approximately 600 ps. As can be seen, the baseline perturbations damp out rapidly.

The RFPG was also designed to be used as a transfer standard for time interval measurements. To accomplish this, all of the timing functions are obtained by digitally counting discrete clock cycles of an internal 10-MHz crystal oscillator. ECL integrated circuits are used exclusively to maintain low timing jitter. The main output pulse and the trigger output pulse come from identical output stages and have identical waveforms. The repetition rate can be set to 1 MHz, 100 kHz, 10 kHz, or 1 kHz. The duty cycle is fixed at 10 percent for the main pulse and 50 percent for the trigger pulse. The trigger-to-pulse delay is selectable at either 0 ns, 100 ns, 200 ns, or 300 ns. The jitter in the delay is 25 ps or less. A mode control sets the outputs to either high dc (1.000 V), low dc (0.000 V), or pulse. The RFPG is built as a single wide plug-in to be operated in a Tek TM-500[1] main frame.

The RFPG is calibrated by using primary lab standards of dc voltage, resistance, and frequency. The transition shape and baseline perturbations of the waveform are calibrated using appropriate measurements, modeling, and computer simulation. It is most important to determine the node voltage $V1$ as a function of time, Fig. 4, and the effects of the nonideal diode
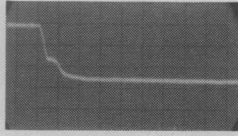
Fig. 7. Driving node $V1$ waveform. Vertical scale is 500 mV/div (center line is 0 V). Horizontal scale is 2 ns/div observed through a 500-Ω, 10:1 probe.

$CR1$ in disconnecting $V1$ from the output. $V1$ can be determined by either time domain measurements, Fig. 7, or computer simulation of the output stage. The major diode effects that must be considered are: 1) nonlinear I–V characteristic; 2) nonlinear C–V characteristic; and 3) charge storage–reverse recovery transient effects. The charge storage effects are minimized by using a high-speed Schottky diode. The output baseline perturbations are primarily due to the driving node waveform $V1$ being capacitively coupled to the output via the reverse-biased diode $CR1$.

Modeling and Computer Simulation

The motivation for doing computer analysis and simulation of the RFPG is that in order to have a waveform generator standard, one needs a theoretical basis for how the electronics generate the various electrical waveforms. Circuit modeling, using known or measured parameters for the circuit model, allows one to approximate the actual results and thereby predict the circuit performance from basic principles, which in turn, yields the available waveform.

The circuit analysis programs developed for the RFPG were written using super-SCEPTRE. Waveform generation by computer simulation of the RFPG was conveniently obtained by accessing via a graphics terminal over a multiplexed hardwired link at 9600 baud.

For the purpose of modeling the RFPG, the ability to utilize time-dependent, nonlinear circuit elements provided for in SCEPTRE was essential [6]. In particular, primary dependent current sources for ideal diodes proved to be especially convenient when combined with voltage- or current-dependent capacitors. Secondary dependent current sources with nonlinear, table-entered dependencies permitted the development of an accurate model for the RF-switching transistors used in the output stage of the RFPG.

The output stage of the RFPG consists of positive and negative current sources that are steered by means of transistor switches. An initial model that was used for simulating the output stage is shown in Fig. 8. The output diode ($CR1$) can be simulated by ideal diode current $JD1$ and associated components $C1$, $L1$, $C4$ and, together with load resistor $R3$, form the output branch of the generator circuit. Similarly, $JD2$, $C2$, $L2$, and $C5$ represent the reverse clamping diode ($CR2$) connected to driving node 4. $C1$ and $C2$ are voltage-dependent junction capacitances, $L1$ and $L2$ are package inductances, and $C4$ and $C5$ are the package capacitances of the respective diodes. $R2$ and $C3$ are the filter components for controlling part of the undershoot contained in the negative going transition of the output waveform. $J1$ and $J2$ are the nominal

20-mA and 30-mA positive and negative constant current sources. To simulate the action of the switching transistors, time-dependent resistors $R1T$ and $R2T$ are utilized. The resultant driving node and output voltage waveforms obtained from this initial model reveal that the actual step-like output waveform is limited by the speed of the transistor switches more than by the properties of the output current-steering diodes.

Fig. 9 shows a simple common-emitter equivalent circuit that was developed for the RF switching transistors, based on a modified Geller, Mantek, and Boyle (GMB) model [7]. The primary difficulty overcome with the use of this model is the effect of the nonlinear base-to-emitter characteristics of the transistor when driven with large signals having low source impedances. The basic GMB model represents the base-emitter junction (from charge control theory) as the nonlinear forward resistance of a diode in parallel with a current-dependent capacitor. An analysis which supports this modified GMB model can be made by noting that for the forward-biased base–emitter junction

$$I_{b'e} = I_0 \left[-1 + ex \frac{e}{kT} (V_{b'e} - E_3) \right] \qquad (3)$$

so that

$$[r_{b'e}]^1 = \frac{dI_{b'e}}{dV_{b'e}} = \frac{e}{kT} (I_{b'e} + I_0) \qquad (4)$$

where $e/kT = 38.7$ at $T = 300$ K, I_0 is the base–emitter reverse saturation current, and E_3 is an offset voltage in the base–emitter I–V characteristic.

It is well known that when the junction transition is switched by means of a large step of base current but does not saturate, the collector current is approximately an exponential function of time [8]. The nonlinear GMB model shown in Fig. 9 simulates this type of response by means of the time constant, $T_{b'} = r_{b'e} C_1$, which is considered to be invariant over the entire base current excursion. Consequently, C_1 is seen to be a function of the nonlinear base–emitter current, since

$$C_1 = \frac{T_{b'}}{r_{b',e}} = \frac{eT_{b'}}{kT} (I_{b'e} + I_0). \qquad (5)$$

The general expression for the static base–emitter I–V characteristic is given by

$$V_{be} = I_b r_{bb'} + V_{D1} + E_3 \qquad (6)$$

where

$$V_{D1} = \frac{kT}{e} \ln \left(\frac{I_b}{I_0} + 1 \right) \qquad (7)$$

is the ideal diode voltage, and the J_1 current source (leakage current of the back-biased collector–base junction) is neglected. Consequently, the bulk resistance $r_{bb'}$ can be found by rearranging (6) and using (7) to give

$$R_1 = r_{bb'} = \frac{1}{I_b} \left[V_{be} - E_3 - \frac{kT}{e} \ln \left(\frac{I_b}{I_0} + 1 \right) \right]. \qquad (8)$$
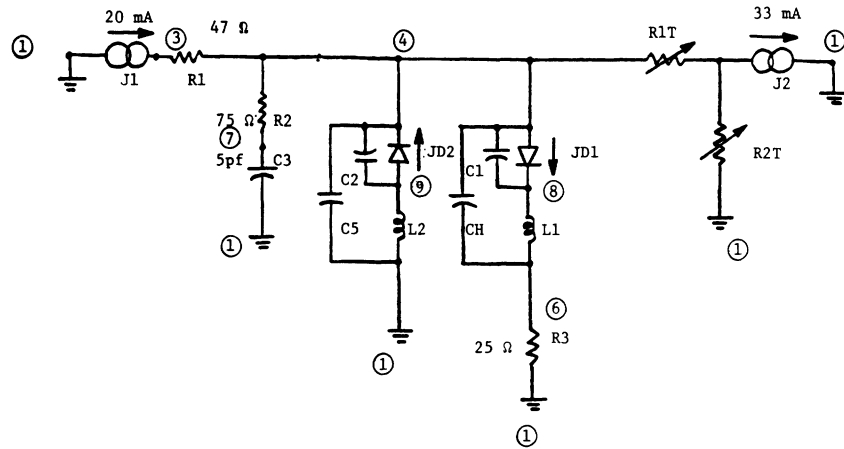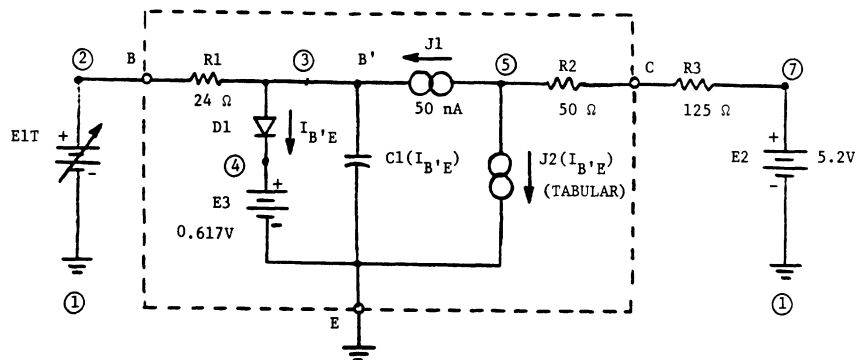
Fig. 8.   An initial RFPG program model.

Fig. 9.   Common-emitter equivalent circuit (modified GMB model) for switching transistors.
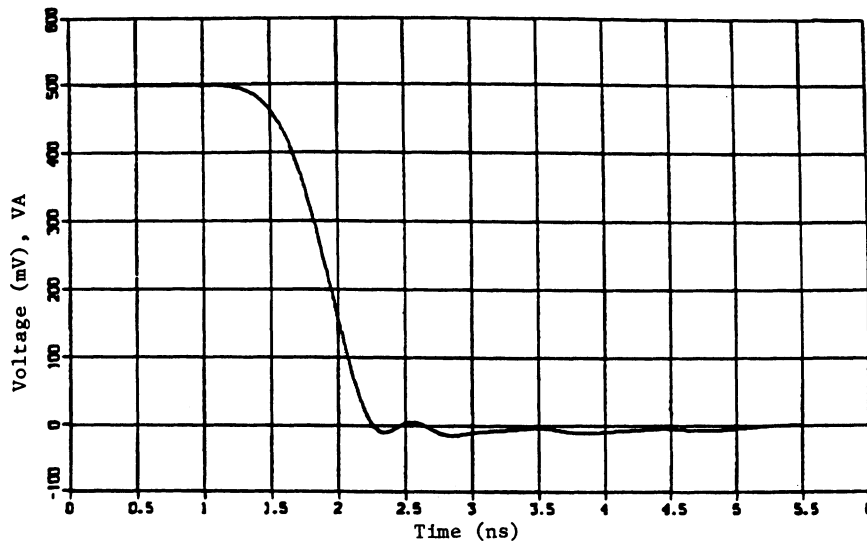

Fig. 10.   Example of the available transition waveform of RFPG across a 50-Ω load (c.f. Fig. 6).

The remaining elements of the model are the conventional hybrid circuit components, except that $J2$ is not simply beta (or α_{fe}) times the base current. Rather, the beta variation with current level is accounted for by means of tabular data. R_2 is the saturation reistance of the collector characteristic.

Finally, the modified GMB model for the transistors, to-gether with appropriate ECL input drive signals, is used to replace the $R1T$-$R2T$ resistors in the basic output stage model of Fig. 8. The use of a more complete model is then possible for investigating the effects of parameter variations on the RFPG output waveform simulation. Fig. 10 shows an example of the available transition waveform across a 50-Ω load of the
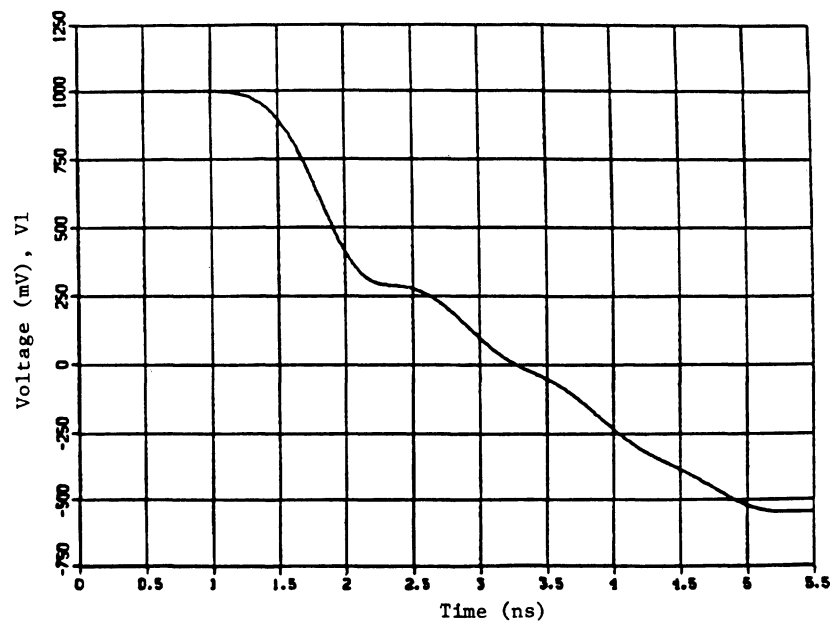
Fig. 11.   Example of corresponding driving node voltage of RFPG (c.f. Fig. 7).

RFPG produced by computer simulation (c.f. Fig. 6). Fig. 11 shows a similar example of the corresponding driving node voltage $V1$ obtained (c.f. Fig. 7) [9].

REFERENCES

[1] W. D. McCaa, Jr. & N. S. Nahman, "Generation of reference waveforms by uniform lossy transmission lines," *IEEE Trans. Instrum. Meas.*, vol. IM-19, no. 4, pp. 382–390, Nov. 1970.

[2] J. R. Andrews, "Picosecond pulse generators using microminiature Mercury switches," National Bureau of Standards, Boulder, CO, Tech. Rep. NBSIR 74-377 Mar. 1974.

[3] G. J. Frye, "A predictable subnanosecond pulse generator," *IEEE Trans. Instrum. Meas.*, vol. IM-17, no. 4, pp. 324–329, Dec. 1968.

[4] N. S. Nahman, R. M. Jickling, and D. R. Holt, "Reference-waveform generation using Debye dielectric dispersion," National Bureau of Standards, Boulder, Co., Tech. Rep. NBSIR 73-704 Dec. 1972, 91 pages (available from NTIS).

[5] J. R. Andrews, "Pulse reference waveform standards development at NBS," in *Proc. 8th ATE Sem.* (Pasadena, CA, Jan. 1981), Boston, MA: Benwill, pp. IV-13-19.

[6] D. Becker, *Extended SCEPTRE*, vol. I, (User's Manual), GTE Sylvania, Dec. 1974.

[7] S. B. Geller, P. A. Mantek, and D. R. Boyle, "A general junction-transistor equivalent circuit for use in large-signal switching analysis," *IRE Trans. Electron. Comput.*, pp. 670–679, Dec. 1961.

[8] J. J. Ebers and J. L. Moll, "Large-signal behavior of junction transistors," *Proc. IRE*, vol. 42, pp. 1761-1772, Dec. 1954.

[9] J. R. Andrews, B. A. Bell, and E. E. Baldwin, "Reference flat pulse generator," NBSIR 82-, to be published.