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Self-Switching Damping Circuit for Reducing Transmitter Ringdown Time in High Power Pulse NMR*

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In this paper we describe a circuit for reducing the transmitter ringdown time and thereby improving the recovery time in pulse NMR experiments. The circuit uses only solid state devices and requires no external switching. For transmitter voltages less than 0.5 V peak to peak the effective resistance in parallel with the transmitter coil in a crossed coil spectrometer is only 6 Ω, whereas for larger voltages the shunting resistance is of the order of 1–2 kΩ. This circuit thus has the effect of significantly squaring the envelope of an rf pulse. It has the extra advantage of suppressing noise generated by the transmitter during the time interval between pulses.

INTRODUCTION

A MAJOR problem in pulse NMR is concerned with the need for obtaining a large rf field over the sample dimensions and then detecting a weak nuclear free induction decay immediately afterwards. Often the rf pulse is of the order of several kilovolts whereas the receiver is sensitive to microvolt signals. In order for us to observe a signal, the rf field must first decay to a value less than the nuclear signal. It takes in excess of 20 time constants for the radio frequency voltage to drop nine orders of magnitude from 1 kV to 1 μV. The time required for it to drop one time constant is given by

$$\tau = 2Q/\omega, \tag{1}$$

since the number of ringing cycles[1] is Q/π. Thus, one way to control the transmitter ringdown in a crossed coil rig is to keep the coil Q low at the sacrifice of higher possible H_1 fields. We feel that it is more desirable to design as efficient a system as possible, and then to reduce the Q by nonlinear loading at the transmitter output. There have been circuits such as one developed by Clark[2] and modified by Spokas[3] which feature vacuum diodes biased in such a way that they appear as a high impedance for large signals, and a much reduced impedance for small signals. The maximum shunting effect possible in this arrangement is the vacuum diode forward biased dynamic resistance of around 250 Ω. We wish to describe a nonlinear shunting circuit using a balanced silicon high speed self-switching diode bridge which results in a shunting resistance of only 6 Ω.

I. CIRCUIT OPERATION

The circuit is shown in Fig. 1. The diode bridge is biased such that the diodes are all forward conducting in the absence of any signal from the power amplifier. Thus with zero voltage V_{rf} from the power amplifier, $\frac{1}{2}I_b$ flows in each diode D_1–D_4. The dynamic resistance R_A seen at A, the junction of D_1 and D_2, is equal to the resistance of the series parallel combination of all four diodes which is numerically the same as r, the resistance of each diode. The current–voltage relation for a silicon junction is given approximately by

$$I = I_0[\exp(V/2V_T) - 1], \tag{2}$$

where I_0 is the leakage current and V_T is about 26 mV[1] at room temperature. Then

$$r = dV/dI = 2V_T/(I+I_0) \cong 2V_T/I \tag{3}$$

for low leakage diodes. Since the bias current per diode, I, is equal to $\frac{1}{2}I_b$, we obtain

$$r = 4V_T/I_b \cong 8V_T R_b/V_b, \tag{4}$$

since $I_b \cong V_b/2R_b$.

When the peak-to-peak rf voltage at A, V_{rf}, exceeds the cut-in voltage of the diodes (approximately 0.5 V), then two of the four diodes in the bridge become back biased and the effective shunt resistance is increased to $2R_b$. To see this, suppose that V_{rf} is present and on a positive excursion. The bias currents of D_1 and D_4 will increase while those of D_2 and D_3 will decrease by the same amount. When the current $\frac{1}{2}I_{rf}$ equals the bias current $\frac{1}{2}I_b$ in the opposite direction, D_2 and D_3 become back biased and I_{rf}

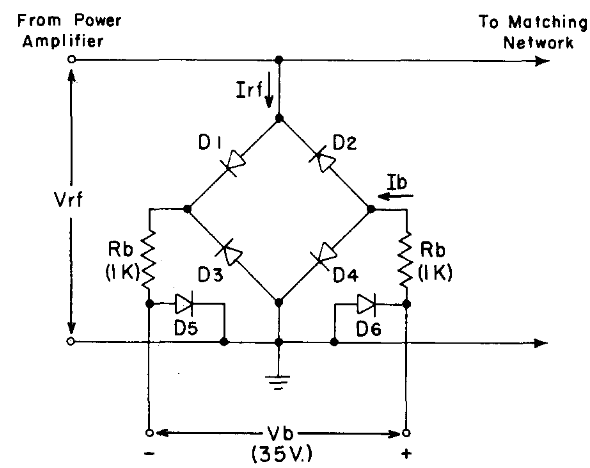

FIG. 1. Schematic circuit of the nonlinear diode bridge shunt. D_1–D_4—VR400X/F/S (bridge); D_5, D_6—GE-504A (600 V, 1 A).

is forced to flow to ground through the series combination of D_1, the two resistors R_b, the supply V_b, and D_4. When $V_{rf}=V_b$, then the clamping diode D_5 becomes forward biased and I_{rf} is shunted to ground there. The shunting resistance then drops from $2R_b$ to R_b. For V_{rf} negative, the roles of the diodes are reversed and the analysis is the same.

We can determine the maximum and minimum shunt values. For low values of V_{rf} (less than 0.5 V peak to peak), the effective resistance R_A between A and ground is just r and is determined by the ratio of R_b and V_b as indicated in Eq. (4). For high V_{rf} values (of peak-to-peak value greater than $2V_b$), the effective resistance at A equals R_b. Our choice of R_b is a compromise between the goal that our power supply approximate a constant current source (large R_b) and the need to get maximum bias current through the diodes (small R_b). For illustration, consider a V_b of 35 V[4] and an R_b of 1000 Ω. Then, for small signals, we can substitute into Eq. (4) to obtain $R_A=r=6\ \Omega$. This is a factor of 42 better than the corresponding shunt resistance for the previous vacuum diode circuits mentioned.[2,3] This reduced shunt resistance has the effect of reducing the ringdown time over that achievable with vacuum tube circuits by the same factor.

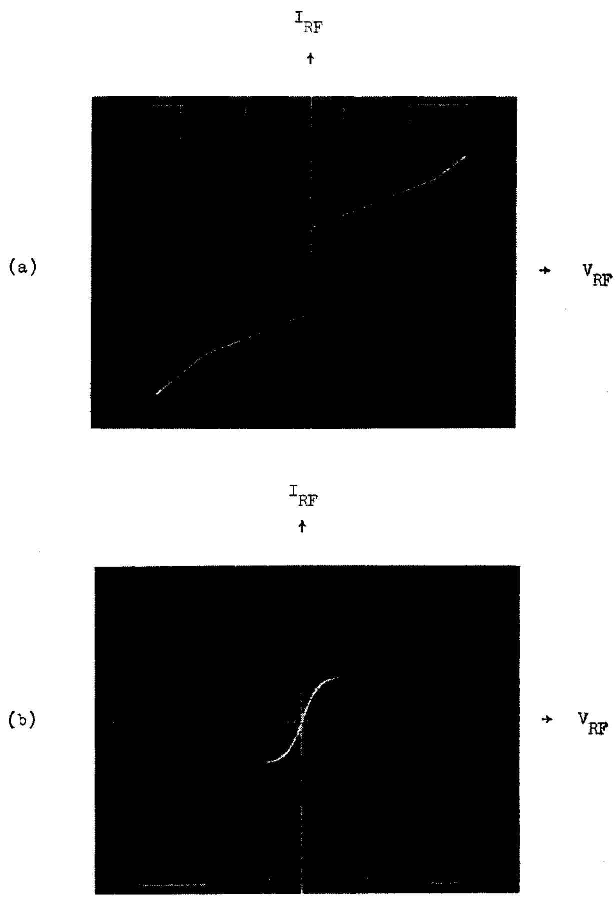

FIG. 2. (a) The dynamic response of the diode bridge shunt. Vertical, 10 mA/div; horizontal, 10 V/div. (b) Same as (a) but horizontal, 0.2 V/div. The measured low voltage resistance of 7 Ω is in excellent agreement with our estimate using Eq. (4).

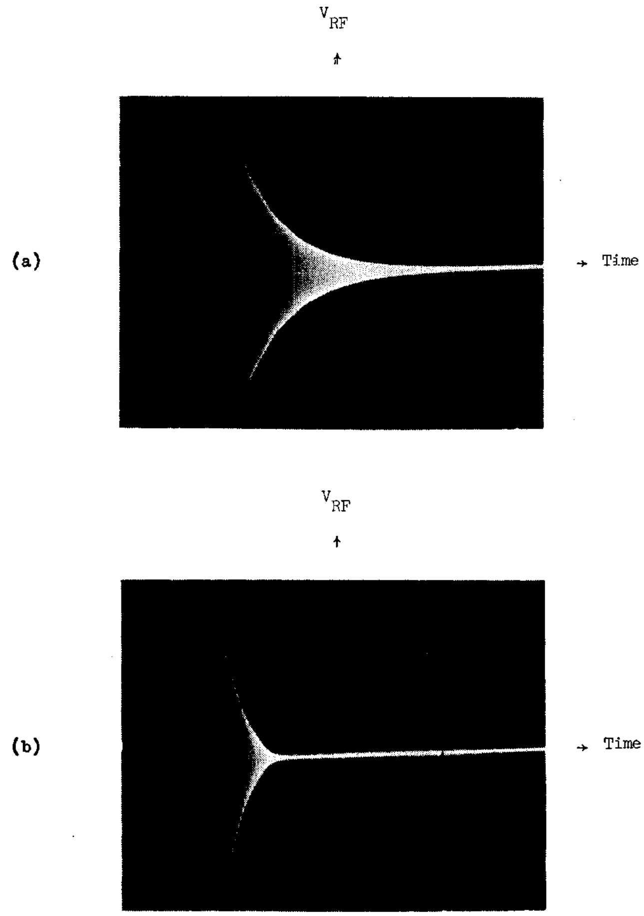

FIG. 3. 4 MHz rf voltage at H_1 coils with diode bridge connected at 50 Ω input to matching circuit. $R_b=1000\ \Omega$. (a) $V_b=0$ V. Vertical, 5 V/div; horizontal, 5 μsec/div. (b) $V_b=35$ V, same scales.

II. PERFORMANCE

Figure 2 represents the characteristics of the actual circuit as observed on a Tektronix 575 curve tracer. The dynamic resistance is represented by the reciprocal of the slope of the curve (smaller slope → increased resistance). Note the increased slope in Fig. 2 (a) which occurs when $|V_{rf}|>|V_b|$ as discussed earlier. The observations are in very good quantitative agreement with the analysis given above.

We have taken advantage of the small value of the circuit impedance at low voltages by placing our shunt bridge at the point where the 50 Ω coaxial output from the transmitter is connected to the matching circuit feeding our H_1 coils. The voltage at this point is well within the breakdown rating of the diodes even at relatively high power levels.[5]

This circuit has resulted in greatly improved transmitter ringdown times as is illustrated in Fig. 3. An additional benefit is that shot noise generated by the transmitter is completely suppressed between pulses.[6]

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[1] J. Millman and H. Taub, *Pulse, Digital, and Switching Waveforms* (McGraw-Hill, New York, 1965), pp. 60, 179.
[2] W. Gilbert Clark, Rev. Sci. Instrum. **35**, 316 (1964).
[3] J. J. Spokas, Rev. Sci. Instrum. **36**, 1436 (1965).
[4] V_b need not be regulated but must be floating since neither terminal is grounded. We used a simple bridge rectifier and filter for this purpose.
[5] The diodes which we used have a peak inverse voltage rating of 400 V, a 10 A current rating, and a reverse recovery time t_{rr} of 200 nsec. The rf voltage across the circuit is only 500 V peak to peak and is stepped up later in the matching network. The diodes in the bridge could be replaced by more in series if it is necessary to employ larger rf voltages. If the bridge is operated with these diodes at frequencies much higher than 5 or 6 MHz more of the rf power will be developed in harmonics of the fundamental frequency with a resulting decrease in the rf power at the fundamental frequency and thus in H_1. Preliminary tests indicate that this situation does not become intolerable at frequencies up to 16 MHz. An additional difficulty results from the fact that at the higher frequencies more power would then be dissipated in the diodes; thus, there is a greater danger that their power rating would be exceeded. For operation at frequencies above 16 MHz, it would probably be necessary to replace these diodes with faster diodes, which may then have to be stacked in series and/or parallel in order to achieve satisfactory power and voltage capabilities.
[6] In our case the noise spikes generated by our Collins power amplifier previously were three or four times larger than the thermal noise produced in our receiver coil. The use of the damping circuit completely suppressed these spikes.

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Oscillating Superleak Second Sound Transducers*

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The construction and performance of mechanical transducers suitable for generating and detecting second sound in pure ^{4}He, and also in dilute ^{3}He–^{4}He mixtures at millidegree temperatures, are described. It is shown that when the normal fluid density is small, the behavior of the transducers can be explained by a simple acoustical model which enables their sensitivity, frequency response, and reflection coefficient to be calculated in terms of easily measured parameters.

INTRODUCTION

As is well known, second sound is simply a density, or first sound, wave in the quasiparticle or excitation gas of a system. In the two liquid systems in which second sound is known to propagate, superfluid ^{4}He and mixtures of ^{3}He in superfluid ^{4}He, an obvious way of generating second sound is by placing a vibrating superleak piston or diaphragm in the liquid. If the superleak is ideal it will modulate the density of the excitation gas, or normal fluid, alone and this density modulation will propagate as a second sound wave. The second sound can also be detected by using a microphone with a superleak for the diaphragm. In this paper we describe an electrostatic transducer using this principle which may be used as both a loudspeaker and a microphone.

The main advantage of this type of transducer over the more usual heater–thermometer arrangement is in generating and detecting second sound in dilute mixtures of ^{3}He in ^{4}He at temperatures below 0.6 K. In this regime the normal fluid is predominantly ^{3}He quasiparticles and second sound is a wave of ^{3}He number density. Thus, although temperature fluctuations are still associated with the wave (the ^{3}He number density changes are adiabatic[1]), at low temperatures a heater mainly produces second sound indirectly via the interaction of thermally generated phonons and rotons with the ^{3}He quasiparticles. The heater method has been used at temperatures down to 0.2 K and ^{3}He concentrations as low as 0.35%,[2] but for the above reasons it is not suitable at lower temperatures. In addition, for measurements in the millidegree temperature range the problems of attaining and controlling the temperature of the sample would be severely aggravated by any large local heating.

Superleak transducers of the type to be described have been successfully used to generate and detect second sound in ^{3}He–^{4}He mixtures with ^{3}He concentrations as low as 0.06% and at temperatures down to 0.03 K,[3] and also in pure ^{4}He at temperatures very close to the λ point.[4]

They would also be suitable for the generation and detection of collisionless or zero second sound at extremely low temperatures in mixtures. In this regime the quasiparticle mean free path is large compared to the wavelength. The possible existence of collisionless second sound has been discussed by a number of authors[5] but no experiments have been carried out yet.

In an appendix to this paper the theoretical efficiency of the oscillating superleak transducer is compared with that of the Peshkov transducer[6] where a solid piston vibrates behind a fixed superleak filter. It is shown that the present system is a more efficient generator when the normal fluid density is small.

I. DESCRIPTION OF TRANSDUCER

A typical transducer unit, consisting of two nominally identical transducers at opposite ends of a cylindrical propagation tube, is shown in Fig. 1. The principle em-