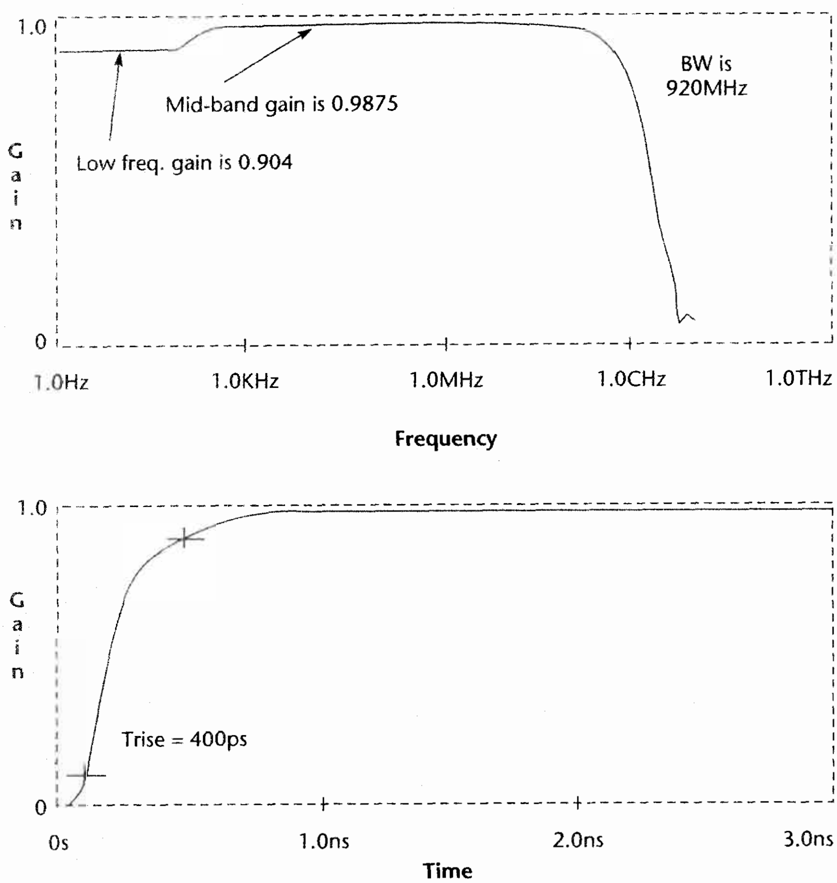


Figure 7–12.
The magnitude and step responses of the bootstrapped source follower.

back gives the two-path source follower the accuracy of a precision op amp. At high frequencies, the signal feeding through C1 dominates control of gate 1, and the source follower operates open loop. The FET is protected by the diodes and the current limiting effects of C1. The 1/f noise of the FET is partially controlled by the op amp, and the circuit can offset large DC levels at the input with the offset control point shown in Figure 7–13.

Figure 7–14 shows the flatness details of the two-path impedance converter. Feedback around the op amp has taken care of the low-frequency gain error exhibited by the bootstrapped source follower (Figure 7–12). The gain is flat from DC to 80MHz to less than 0.1%. The "wiggle" in the magnitude response occurs where the low- and high-frequency paths cross over.

There are additional benefits to the two-path approach. It allows us to design the high-frequency path through C1 and the MOSFET without regard to DC accuracy. The DC level of the impedance converter output is independent of the input and can be tailored to the needs of the preamplifier. Although it is not shown in the figures, AC coupling is easily implemented by blocking DC to the non-inverting input of the op amp.

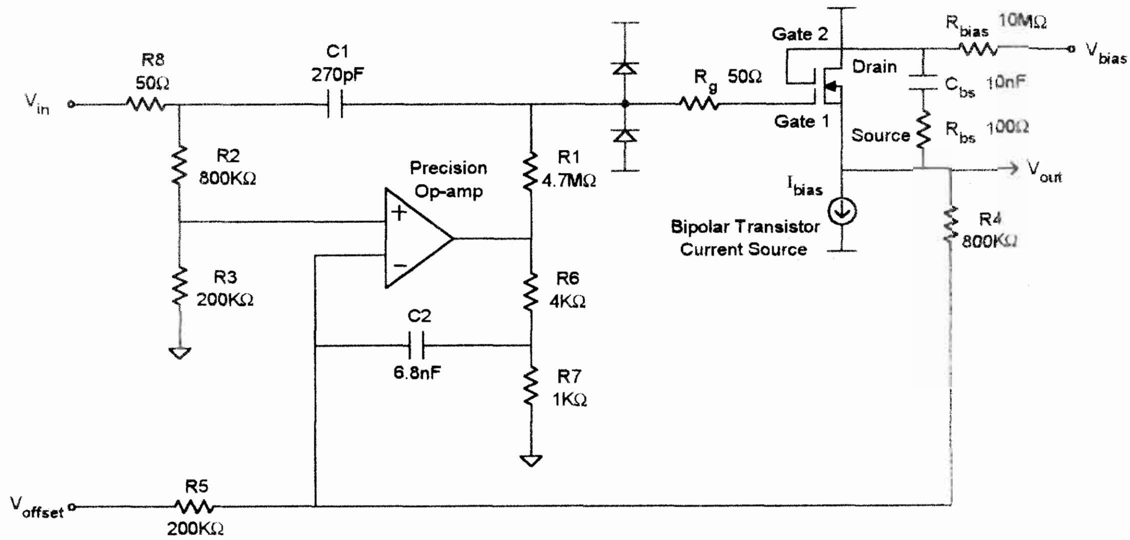


Figure 7–13. A two-path impedance converter.

Thus we avoid putting an AC coupling relay, with all its parasitic effects, in the high-frequency path.

There are drawbacks to the two-path impedance converter. The small flatness errors shown in Figure 7–14 never seem to go away, regardless of the many alternative two-path architectures we try. Also, C1 forms a capacitive voltage divider with the input capacitance of the source follower. Along with the fact that the source follower gain is less than unity, this means that the gain of the low-frequency path may not match that of the high-frequency path. Component variations cause the flatness to vary further. Since the impedance converter is driven by a precision high-impedance attenuator, it must have a very well-behaved input impedance that closely resembles a simple RC parallel circuit. In this regard the most common problem occurs when the op amp has insufficient speed and fails to bootstrap R1 in Figure 7–13 to high enough frequencies.

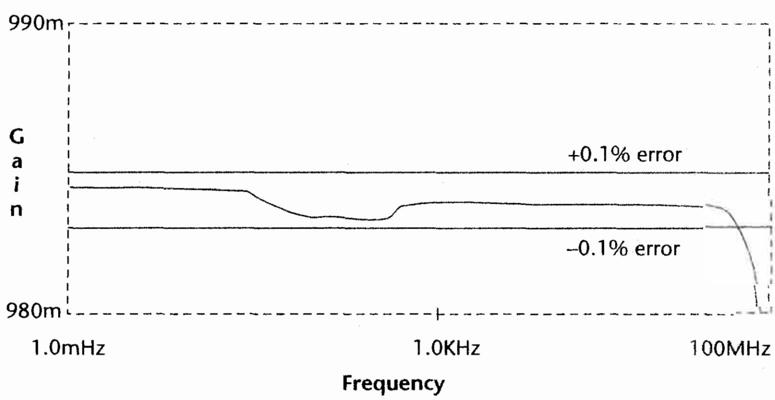


Figure 7–14. Flatness details of the two-path impedance converter.

The overdrive recovery performance of a two-path amplifier can be abysmal. There are two ways in which overdrive problems occur. If a signal is large enough to turn on one of the protection diodes, C1 charges very quickly through the low impedance of the diode (Figure 7–13). As if it were not bad enough that the input impedance in overdrive looks like 270pF, recovery occurs with a time constant of 270pF • 4.7MΩ, or 1.3ms! Feedback around the op amp actually accelerates recovery somewhat but recovery still takes eons compared to the 400ps rise time! Another overdrive mechanism is saturation of the source follower. When saturation occurs, the op amp integrates the error it sees between the input and source follower output, charging its 6.8nF feedback capacitor. Recovery occurs over milliseconds. The seriousness of these overdrive recovery problems is mitigated by the fact that with careful design it can take approximately ±2V to saturate the MOSFET and ±5V to activate the protection diodes. Thus, to overdrive the system, it takes a signal about ten times the full-scale input range of the pre-amp.

I apologize for turning a simple, elegant, single transistor source follower into the "bootstrapped, two-path impedance converter." But as I stated at the beginning, it is the combination of requirements that drives us to such extremes. It is very hard to meet all the requirements at once with a simple circuit. In the next section, I will extend the two-path technique to the attenuator to great advantage. Perhaps there the two-path method will fully justify its complexity.

The Attenuator

I have expended a large number of words and pictures on the impedance converter, so I will more briefly describe the attenuator. I will confine myself to an introduction to the design and performance issues and then illustrate some interesting alternatives for constructing attenuators. The purpose of the attenuator is to reduce the dynamic range requirements placed on the impedance converter and pre-amp. The attenuator must handle stresses as high as ±400V, as well as electrostatic discharge. The attenuator maintains a 1MΩ input resistance on all ranges and attains microwave bandwidths with excellent flatness. No small-signal microwave semiconductors can survive the high input voltages, so high-frequency oscilloscope attenuators are built with all passive components and electromechanical relays for switches.

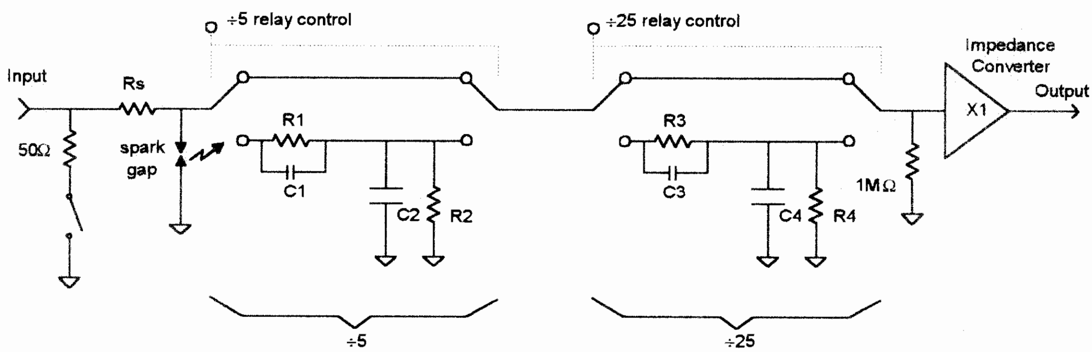
Figure 7–15 is a simplified schematic of a 1MΩ attenuator. It uses two stages of the well-known "compensated voltage divider" circuit. One stage divides by five and the other by 25, so that division ratios of 1, 5, 25, and 125 are possible. There are two key requirements for the attenuator. First, as shown in Figure 7–3, we must maintain $R_1C_1 = R_2C_2$ in the ÷5 stage to achieve a flat frequency response. A similar requirement holds for the ÷25 stage. Second, the input resistance and capacitance at each stage must match those of the impedance converter and remain very nearly constant, independent of the switch positions. This requirement assures that we maintain attenuation accuracy and flatness for all four combinations of attenuator relay settings.

Dividing by a high ratio such as 125 is similar to trying to build a high-isolation switch; the signal attempts to bypass the divider, causing feed-through problems. If we set a standard for feedthrough of less than one least-significant bit in an 8-bit digital oscilloscope, the attenuator must isolate the input from the output by $20\log_{10}(125 \cdot 2^8) = 90\text{dB}$! I once spent two months tracking down such an isolation problem and traced it to wave guide propagation and cavity resonance at 2GHz inside the metallic attenuator cover.

Relays are used for the switches because they have low contact impedance, high isolation, and high withstanding voltages. However, in a realm where 1mm of wire looks like a transmission line, the relays have dreadful parasitics. To make matters worse, the relays are large enough to spread the attenuator out over an area of about 2 × 3cm. Assuming a propagation velocity of half the speed of light, three centimeters takes 200ps, which is dangerously close to the 700ps rise time of a 500MHz oscilloscope. In spite of the fact that I have said we can have no transmission lines in a high-impedance attenuator, we have to deal with them anyway! To deal with transmission line and parasitic reactance effects, a real attenuator includes many termination and damping resistors not shown in Figure 7–15.

Rather than going into extreme detail about the conventional attenuator of Figure 7–15, it would be more interesting to ask if we could somehow eliminate the large and unreliable electromechanical relays. Consider the slightly different implementation of the two-path impedance converter depicted in Figure 7–16. The gate of the depletion MOSFET is self-biased by the 22MΩ resistor so that it operates at zero gate source voltage. If the input and output voltages differ, feedback via the op amp and bipolar current source reduces the error to zero. To understand this circuit, it helps to note that the impedance looking into the source of a self-biased FET is very high. Thus the collector of the bipolar current source sees a

Figure 7–15. A simplified two-stage high-impedance attenuator.



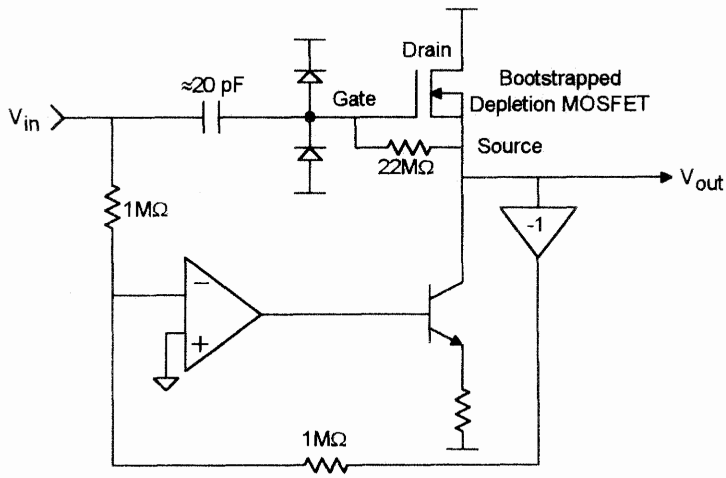


Figure 7–16. A variation on the two-path impedance converter.

high-impedance load. Slight changes in the op amp output can therefore produce significant changes in the circuit output.

The impedance converter of Figure 7–16 can easily be turned into a fixed attenuator, as shown in Figure 7–17. As before, there is a high-frequency and a low-frequency path, but now each divides by ten. There is an analog multiplier in the feedback path to make fine adjustments to the low-frequency gain. The multiplier matches the low- and high-frequency paths to achieve a high degree of flatness. A calibration procedure determines the appropriate gain for the multiplier.

Now we can build a complete two-path attenuator with switched attenuation, as shown in Figure 7–18 (Roach 1992). Instead of cascading attenuator stages, we have arranged them in parallel. In place of the two double-pole double-throw (DPDT) relays of Figure 7–15, we now need only two single-pole single-throw (SPST) relays. Note that there is no need for a switch in the ÷100 path because any signal within range for

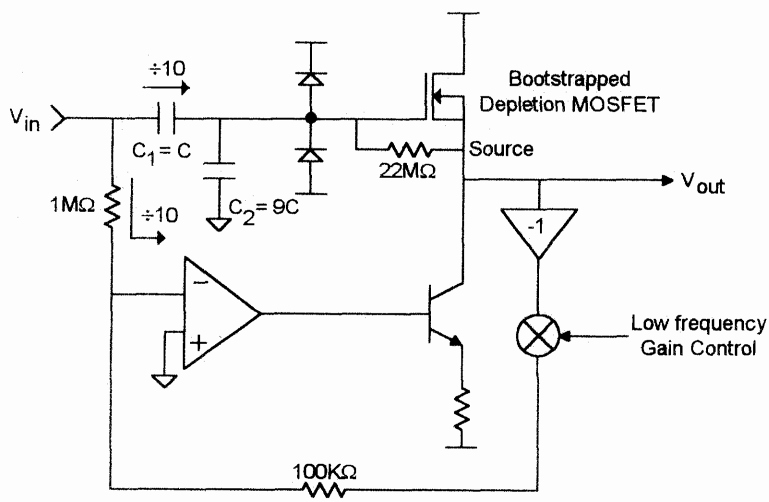


Figure 7–17. An attenuating impedance converter, or "two-path attenuator."

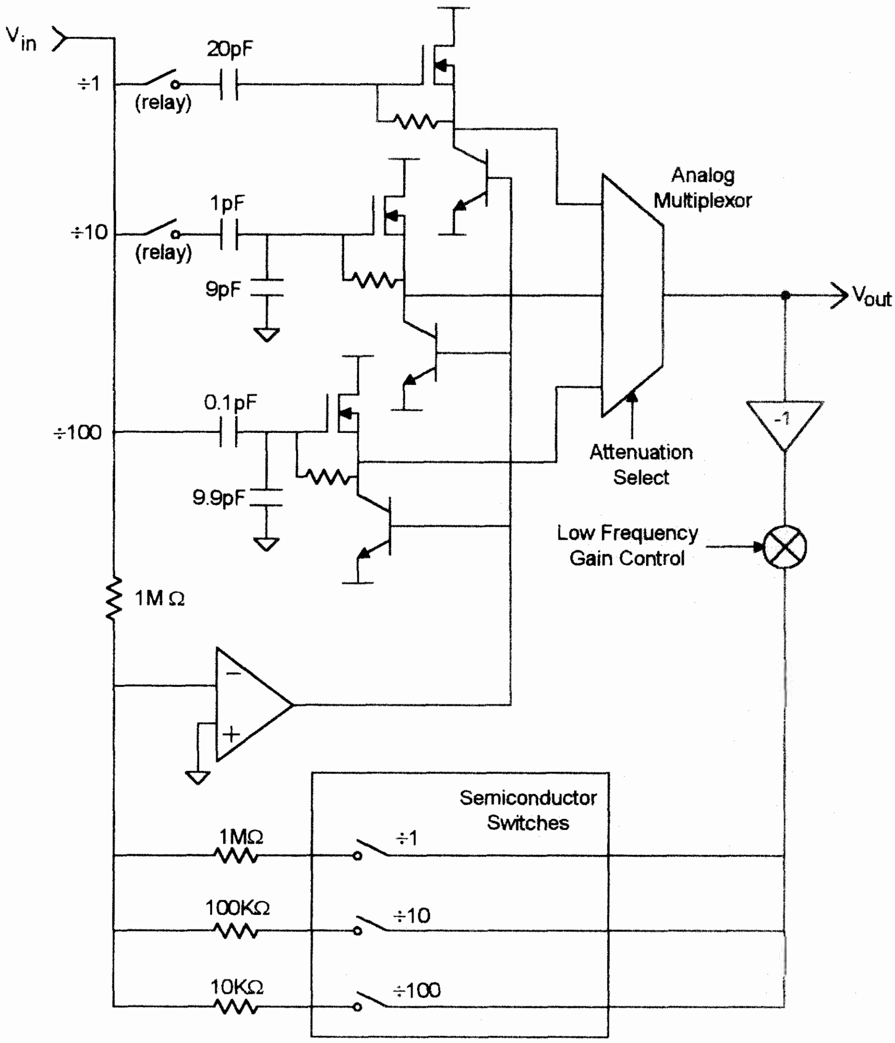


Figure 7–18. A two-path attenuator and impedance converter using only two SPST electromechanical relays. The protection diodes and some resistors are omitted for clarity.

the ÷1 or ÷10 path is automatically in range for the ÷100 path. The switches in the low-frequency feedback path are not exposed to high voltages and therefore can be semiconductor devices.

A number of advantages accrue from the two-path attenuator of Figure 7–18. The SPST relays are simpler than the original relays, and the high-frequency path is entirely AC coupled! The relays could be replaced with capacitive switches, eliminating the reliability problems of DC contacts. One of the most important contributions is that we no longer have to precisely trim passive components as we did in Figure 7–15 to make $R_1C_1 = R_2C_2$. This feature eliminates adjustable capacitors in printed circuit (PC) board attenuators and difficult laser trimming procedures on hybrids. With the need for laser trimming eliminated, we can build on inexpensive PC board attenuators that formerly required expensive hybrids.

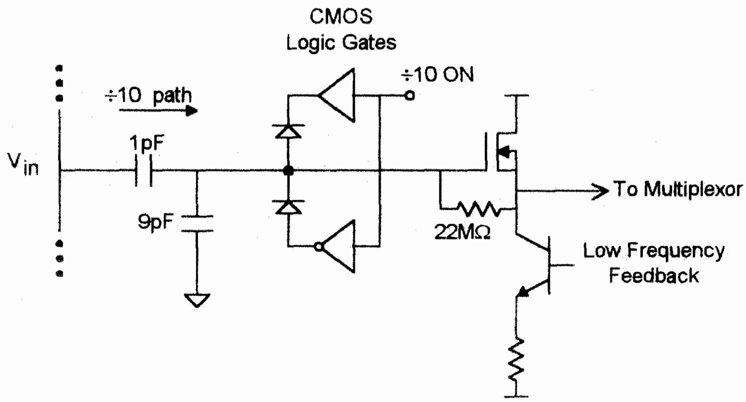


Figure 7–19. Using the protection diodes as switches in the ÷10 path.

We can take the new attenuator configuration of Figure 7–18 further. First observe that we can eliminate the ÷10 relay in Figure 7–18, as shown in Figure 7–19. The diodes are reverse biased to turn the ÷10 path on and forward biased to turn it off. Forward biasing the diodes shorts the 1pF capacitor to ground, thereby shunting the signal and cutting off the ÷10 path. The input capacitance changes by only 0.1pF when we switch the ÷10 path.

Now we are down to one electromechanical relay in the ÷1 path. We can eliminate it by moving the switch from the gate side of the source follower FET to the drain and source, as shown in Figure 7–20. In doing so we have made two switches from one, but that will turn out to be a good trade. With the ÷1 switches closed, the drain and source of the FET are connected to the circuit and the ÷1 path functions in the usual manner. The protection diodes are biased to ±5V to protect the FET.

To cut off the ÷1 path, the drain and source switches are opened, leaving those terminals floating. With the switches open, a voltage change at

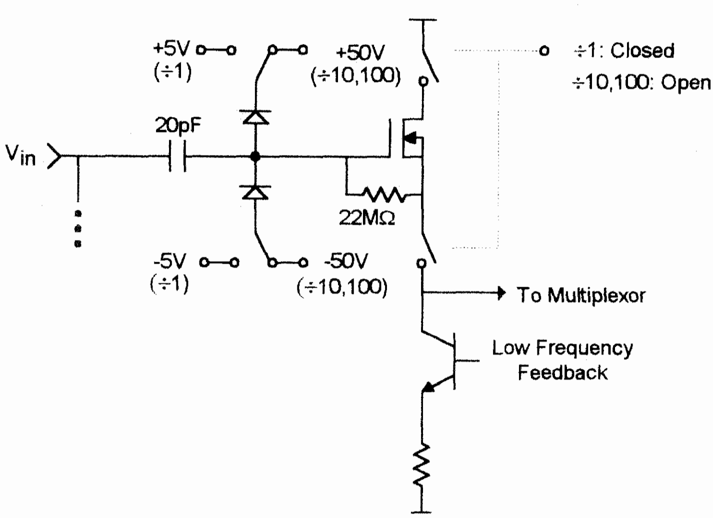
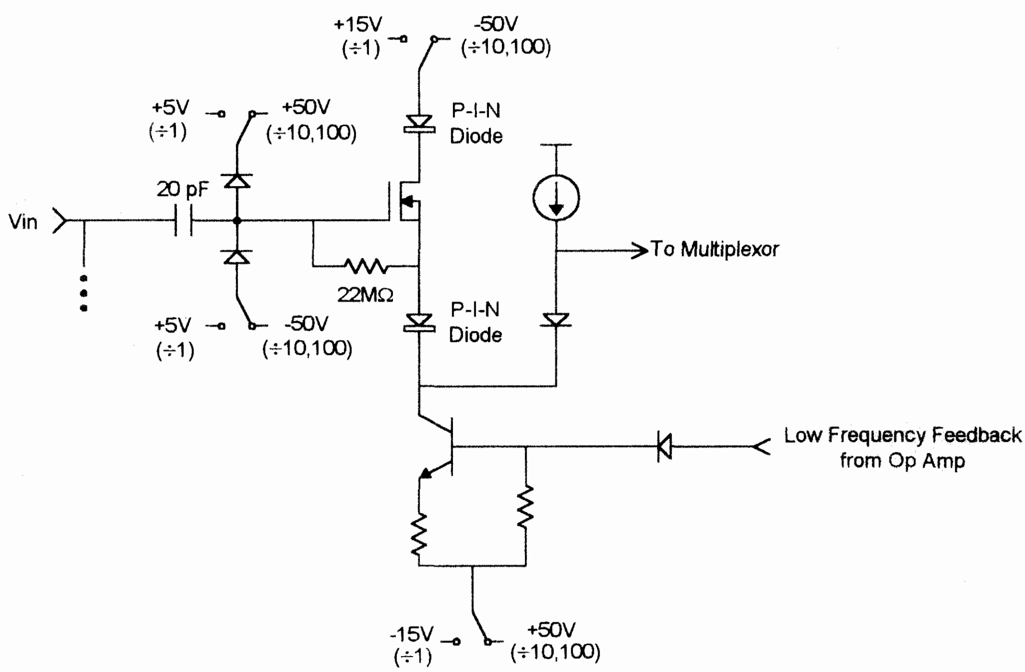


Figure 7–20. Moving the ÷1 switch from the high-impedance input side to the low-impedance output side of the FET.

the input drives the gate, source, and drain of the FET through an equal change via the 20pF input capacitor and the gate-drain and gate-source capacitances. Since all three terminals of the FET remain at the same voltage, the FET is safe from overvoltage stress. Of course, the switches must have very low capacitance in the open state, or capacitive voltage division would allow the terminals of the FET to see differing voltages. In ÷100 mode, the floating FET will see 40V excursions (eight divisions on the oscilloscope screen at 5V per division) as a matter of course. For this reason the ÷1 protection diodes must be switched to a higher bias voltage (±50V) when in the ÷10 and ÷100 modes. The switches that control the voltage on the protection diodes are not involved in the high-frequency performance of the front-end and therefore can be implemented with slow, high-voltage semiconductors.

Can we replace the switches in the drain and source with semiconductor devices? The answer is yes, as Figure 7–21 shows. The relays in the drain and source have been replaced by PIN diodes. PIN diodes are made with a p-type silicon layer (P), an intrinsic or undoped layer (I), and an n-type layer (N). The intrinsic layer is relatively thick, giving the diode high breakdown voltage and extremely low reverse-biased capacitance. A representative packaged PIN diode has 100V reverse breakdown and only 0.08pF junction capacitance. To turn the ÷1 path of Figure 7–21 on, the switches are all set to their "÷1" positions. The PIN diodes are then forward biased, the bipolar transistor is connected to the op amp, and the FET is conducting. To turn the path off, the switches are set to their "÷10,100" positions, reverse-biasing the PIN diodes. Since these switches

Figure 7–21. Using PIN diodes to eliminate the relays in the ÷1 path.



are not involved in the high-frequency signal path, they too can be built with slow, high-voltage semiconductors.

The complete circuit is now too involved to show in one piece on the page of a book, so please use your imagination. We have eliminated all electromechanical switches and have a solid-state oscilloscope front-end. Although I had a great deal of fun inventing this circuit, I do not think it points the direction to future oscilloscope front-ends. Already research is under way on microscopic relays built with semiconductor micro-machining techniques (Hackett 1991). These relays are built on the surface of silicon or gallium arsenide wafers, using photolithography techniques, and measure only 0.5mm in their largest dimension. The contacts open only a few microns, but they maintain high breakdown voltages (100s of volts) because the breakdown voltages of neutral gases are highly nonlinear and not even monotonic for extremely small spacing. The contacts are so small that the inter-contact capacitance in the open state is only a few femtofarads (a femtofarad is 0.001 picofarads). Thus the isolation of the relays is extraordinary! Perhaps best of all, they are electrostatically actuated and consume near zero power. I believe micro-machined relays are a revolution in the wings for oscilloscope front-ends. I eagerly anticipate that they will dramatically improve the performance of analog switches in many applications. Apparently, even a device as old as the electromechanical relay is still fertile ground for a few ambitious inventors!

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